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PARAMETER INDEPENDENT DESIGN UTILIZING SCATTERING PARAMETERS.(U)

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UTILIZING SCATTERING PARAMETERS

THESIS

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GE/EE/77-5

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PARAMETER INDEPENDENT DESIGN UTILIZING SCATTERING PARAMETERS

THESIS

Presented to the Faculty of the School of Engineering
of the Air Force Institute of Technology
Air University
in Partial Fulfillment of the
Requirements for the Degree of
Master of Science

by

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June 1977

Preface

This thesis is an extension at the graduate level of my interest in parameter independent design that was first encountered in an under-graduate transistor amplifier design course, taught at Oklahoma State University.

The reader is assumed to have a basic familiarity with amplifier design and linear algebra; otherwise the content is self-contained. If more detailed theory is desired, the bibliography provides a path directly to most of the major authors.

I wish to acknowledge my deep indebtedness to my very patient thesis advisor Dr. William Davis, my expert typist Mrs. E. L. Davis, and to my wife who knows the art of gently pushing without becoming exasperated.

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Abstract

A parameter independence factor for two-port networks is defined and a method for its calculation using finite differences is shown. An approach to two-port scattering-parameter circuit design using computer optimization techniques is developed and illustrations are presented to demonstrate the utilization of a digital computer for implementing this approach. The versatility of the approach is clarified by demonstrating how both standard network design criteria and parameter independent network design requirements are specified and met. This design technique has direct Air Force application in the areas of microwave network design and Electronic Warfare with particular emphasis on the independence of the network parameters with respect to device parameters.

I. Introduction

Background

The first generation of transistors suffered from the two major drawbacks of expense and silicon impurities causing an uncertainty in the characteristics of otherwise identical transistors. This uncertainty in the parameters caused early transistor circuit design to be more of an art form than an engineering discipline. Due to the low gain and high cost, most circuits were designed to obtain as much gain as possible by providing many custom adjustable components. Consequently early transistorized circuits had to be realigned or trimmed each time a transistor was replaced to insure the desired characteristics.

As the gain of the average transistor increased and the price declined the designer found it no longer necessary to design for near minimum gain, but could concentrate on developing design techniques to minimize the still troublesome spread in individual parameters. By utilizing only part of the total available current gain of a transistor the designer was able to achieve a total circuit gain that could easily be met by transistors having a wider variation in individual gain. For example, if a single stage current gain of 19 was specified and transistors with current gains of 100 and 50 respectively were tried they would yield similar results.

However, when placed in a circuit designed for a gain of 75, they would yield totally different results.

Two design techniques that were developed which provided lower but more stable amplifier gains are referred to as emmiter degeneration and negative feedback, either of which diverts some of the same output signal back to the amplifier input. Both of these techniques are still widely used at audio and low RF frequencies even though transistor purity has increased to the point of practically eliminating the original parameter spread problem. The reason for their continued use is that they also provide compensation for the variation in gain of a single transistor when the frequency is increased. Therefore, these techniques provide for circuits with an improved constant gain characteristic over a broader frequency range than the individual transistor normally will provide.

As attempts were made to design and produce transistors for operation above 100 MHz it became obvious that the silicon used in producing the transistors required higher purity with increased frequency. Increasing purity is not easily achieved; consequently, there has been a definite lag between the initial requirements of the manufacturer and the availability of silicon of such high purity levels. This lag has caused microwave transistors to experience a similar developmental history as the early transistors already discussed. Microwave transistors are now at that point in

development where their cost is down and their gain is high enough to allow the designer the luxury of not having to utilize all the gain available per transistor.

The parameter spread is still a problem which dictates customized adjustment of each circuit both initially and after each transistor replacement. Field replacement of amplifier assemblies, with the manufacturer doing the individual transistor replacement and alignment, is the normal mode for performing microwave transistor amplifier maintenance and is extremely expensive. If field replacement of the transistors could be made without customized trimming and complicated alignment, the majority of the expense of this replacement could be eliminated.

Field replacement as a design goal can only be realized if either the present spread of parameters is totally eliminated (not presently possible) or if a design technique can be found whereby the effects on the overall amplifier characteristics of the transistor parameter variations can be minimized to a negligible or manageable level (parameter independent design). It is the goal of this thesis to develop such a parameter independent technique.

A review of the two previously discussed methods of achieving parameter independence reveals that neither method is directly applicable at such high frequencies. "Emitter degeneration" is not usable in that the required emitter resistor is usually not a pure resistance at such

frequencies but is a complex reactive circuit which represents severe stability problems for the amplifier often resulting in the amplifier self-oscillating at some undetermined frequency. "Negative feedback" is by definition the feeding back of some output back to the input but 180 deg out of phase with normal input. Therefore, for "negative feedback" to be usable the feedback path must maintain a near constant phase shift of the feedback over the desired frequency range of the amplifier. At audio and low radio frequencies the length of the feedback path was normally negligible in comparison to a wave length. At microwave frequencies the required path length is not normally negligible. As the frequency increases the additional phase shift due to the feedback path length increases. Therefore, the feedback phase is highly frequency dependent and may result in associated oscillations.

An alternate approach to these two methods is to derive some measure of the dependency of the circuit response on the active element (transistor) response and to vary the values of the existing circuit components such as to retain all original desired amplifier characteristics while minimizing the parameter dependence of the circuit. A "parameter independence factor" (PIF) is developed using the overall circuit parameters and the device parameters. At microwave frequencies, the simply and accurately measured scattering parameters (S-parameters) may be used to derive a

parameter independence factor (PIF).

This thesis has the two primary goals of developing a method for calculating a parameter independence factor of a circuit and demonstrating how this factor can be used in circuit design. An empirical approach for calculating a parameter independence factor was first considered and then discarded in favor of a "finite-difference" numerical method as shown in Chapter III.

As a basis for evaluating the overall S-parameters of a circuit and for calculating the parameter independence factor, a computer-aided-design program was developed and is presented for use on a digital computer. This program evaluates the overall scattering parameters and implements the necessary finite difference calculation of the parameter independence factor for a circuit both before and after optimization. This optimization may be done for a combination of circuit characteristics including parameter independence.

The concept of constraining the parameter independence of a circuit will be shown to be feasible. The computer program has been able to supply new optimal component values which provide an improvement in the parameter independence of those circuits used while retaining the original desired characteristics in a large class of problems.

The remainder of this thesis provides further details leading to these conclusions. In Chapter II the history and

theory of S-parameters and S-parameter design are covered with emphasis placed on the utility of these parameters. Chapter III provides a detailed examination of the parameter independence factor, how it is numerically calculated and the error function as it is used by the optimization program. Full coverage is also given to the modeling techniques used by the computer program in describing and interconnecting two-port networks. The results of computer calculations for four basic networks are presented in Chapter V which demonstrate usefulness of this approach to parameter independent design.

II. S-Parameters

Since the late 1950s, the bulk of all state-of-the-art microwave circuit design has centered around applications of scattering parameters (S-parameters). This chapter provides the basic S-parameter theory that is used throughout the remainder of the thesis. The historical development of scattering parameters is presented along with a derivation of these parameters from the hybrid (h) parameters. It will also be shown that each of the S-parameters represents a measurable physical relationship and that through them more meaningful and accurate calculations can be made of microwave network characteristics.

Historical Development and Theory

Scattering parameters were originally developed during World War II (WW II) to describe wave-guide and transmission line systems (Ref 5). Until the middle 1960s, no serious attempt to utilize S-parameters for other applications had ever been made.

Early in the development of transistor design, hybrid parameters (h-parameters) became the accepted parameters for modeling the small signal transistor characteristics. The h-parameters for a two-port network are defined by the following equations

$$\begin{aligned} V_1 &= h_{11}\, I_1 + h_{12}\, V_2 \\ I_2 &= h_{21}\, I_1 + H_{22}\, V_2 \end{aligned} \qquad (1)$$

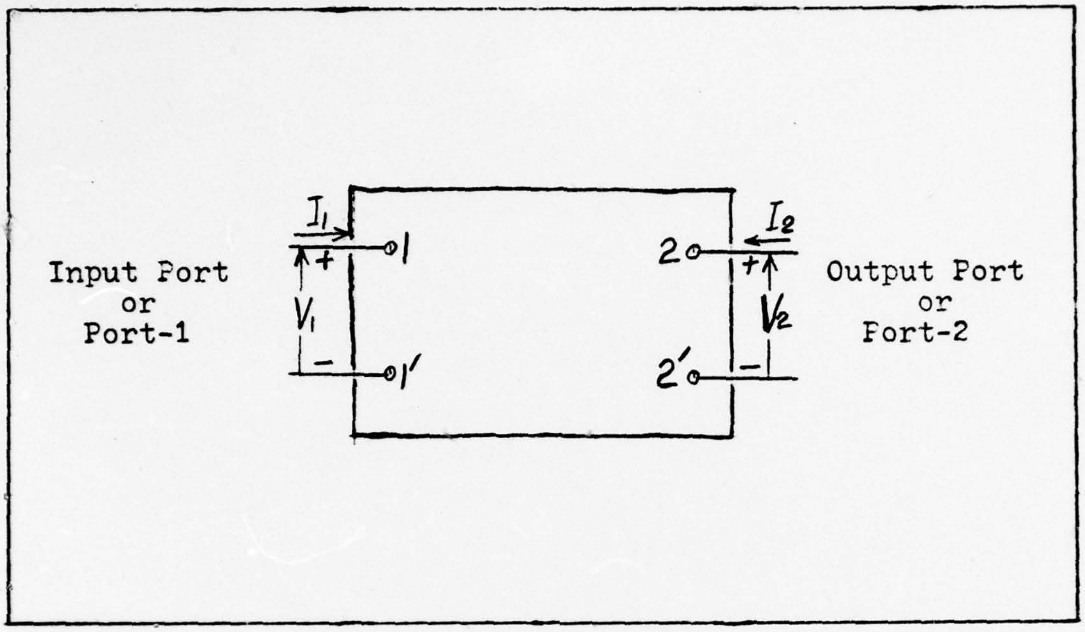

Fig. 1. Two-Port Network

describing the two-port network shown in Fig. 1. The measurement of h_{11} requires V_2 to be zero as by placing an A.C. short circuit across the output terminals of the two-port of Fig. 1. Thus from Eq (1) h_{11} is

$$h_{11} = \left. \frac{V_1}{I_1} \right|_{V_2 = 0} \tag{2}$$

and h_{11} represents the input resistance with the output shorted. The other h-parameters can be measured in a similar manner by terminating either the input with an open circuit or the output with a short circuit.

This requirement of using a short-circuit or an open-circuit load carries with it several disadvantages at high frequencies. First, good short or open circuits are hard to achieve at a single frequency much less over a band of

frequencies. Secondly, active devices (e.g., transistors and tunnel diodes) tend to become short and open circuit unstable as frequency increases. Thus, in many instances the measurement of the h, y, or z-parameters, all of which require open or short circuit loads, is impossible due to device oscillation. Lastly, at frequencies above 30 MHz it becomes increasingly hard to find equipment available which can readily measure the total voltage or current at the ports of a network.

As a consequence of the above disadvantages, highly specialized equipment such as the Transfer-Function and Immittance Bridge of General Radio Company came into use. Even though this meter is capable of measuring the high frequency parameters of a transistor, it still does not provide a reliable solution to the instability problem previously mentioned. In addition, this bridge must be recalibrated at each frequency and complex modifications made in measuring reflection and transmission characteristics. Thus, this meter is too tedious and time consuming to use for broad band measurements.

Above 30 MHz, the higher the frequency the easier it becomes to make distributed or traveling-wave measurements rather than measure discrete variables (i.e., total voltage or total current). Distributed or traveling-wave measurements have been the norm for microwave measurements since before WW II. Thus, due to the problems inherent in measuring conventional parameters z, y, or h, a set of parameters

was needed which did not suffer from the previous disadvantages and which was measurable using distributed measuring techniques. This set of distributed parameters is derivable from the two-port analysis of a transmission line.

Along a transmission line the total voltage V_t and total current I_t at any point can be broken down into an incident (i) and reflected (r) wave components

$$V_t = V_i + V_r$$

$$I_t = \frac{V_i - V_r}{Z_o} \quad . \tag{3}$$

Using this wave concept, the voltage and current at terminals (1-1') and (2-2') of the two-port shown in Fig. 1 are

$$V_1 = V_{1i} + V_{1r} \ , \ V_2 = V_{2i} + V_{2r}$$

$$I_1 = \frac{V_{1i} - V_{1r}}{Z_o} \ , \ I_2 = \frac{V_{2i} - V_{2r}}{Z_o} \quad . \tag{4}$$

The real impedance Z_o for a lossless line is used in these equations, assuming that the reference impedance is the same for both input and output measurements. Substituting Eq (4) into Eq (1) produces a set of parameters relating the incident and reflected voltage waves as

$$V_{1r} = \frac{\Delta h + h_{11} - h_{22} - 1}{\Delta h + h_{11} + h_{22} + 1} V_{1i} + \frac{2h_{12}}{\Delta h + h_{11} + h_{22} + 1} V_{2i}$$

$$V_{2r} = \frac{-2h_{2}}{\Delta h + h_{11} + h_{22} + 1} V_{1i} - \frac{\Delta h - h_{11} + h_{22} - 1}{\Delta h + h_{11} + h_{22} + 1} V_{2i} \tag{5}$$

or

$$V_{1r} = S_{11}V_{1i} + S_{12}V_{2i}$$
$$V_{2r} = S_{21}V_{1i} + S_{22}V_{2i} \qquad (6)$$

where $\Delta h = h_{11}h_{22} - h_{12}h_{21}$ and the parameters S_{kl} are functions of the original h-parameters. If Eqs (6) are divided by $\sqrt{Z_o}$ we have

$$\frac{V_{1r}}{\sqrt{Z_o}} = S_{11}(h)\frac{V_{1i}}{\sqrt{Z_o}} + S_{12}\frac{V_{2i}}{\sqrt{Z_o}}$$
$$\frac{V_{2r}}{\sqrt{Z_o}} = S_{21}(h)\frac{V_{1i}}{\sqrt{Z_o}} + S_{22}\frac{V_{2i}}{\sqrt{Z_o}} \qquad (7)$$

where these S-parameters represent a form of scattering-parameters and the incident and reflected voltages divided by $\sqrt{Z_o}$ represent the square roots of the incident and reflected power respectively.

In 1948 Dicke (Ref 5) proposed and in 1965 Kurokawa (Ref 10) expanded on a concept of incident and reflected power waves a_n and b_n defined as

$$a_n = \frac{V_n + Z_nI_n}{2\sqrt{|ReZ_n|}} \qquad (8)$$

$$b_n = \frac{V_n - Z_n^*I_n}{2\sqrt{|ReZ_n|}} \qquad (9)$$

where n is the n^{th} port of a multiport network. In Eq (8), $V_n + Z_nI_n$ represents a Thevenin-equivalent voltage source for the generator and in Eq (9), $V_n - Z_n^*I_n$ represents a

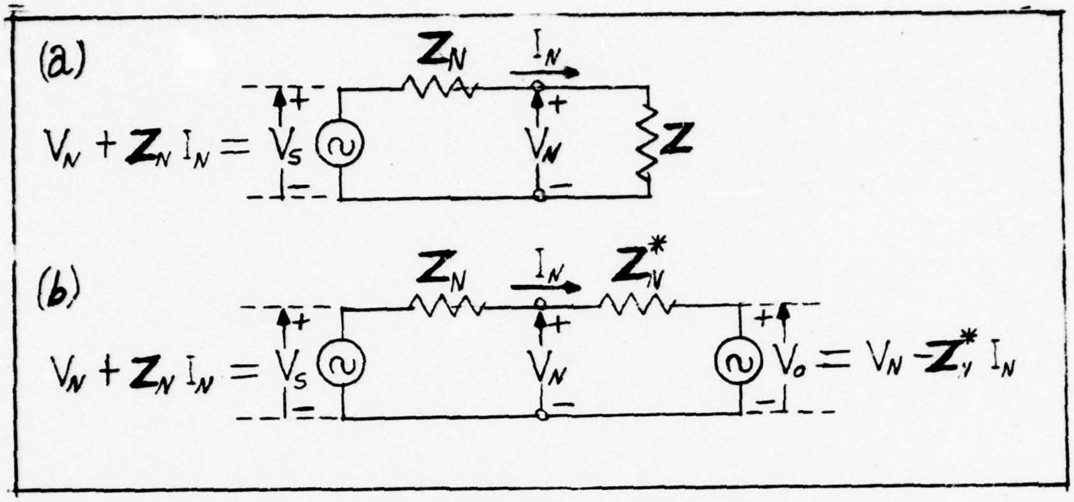

Fig. 2. Thevenin Equivalent Circuit

Thevenin-equivalent dependent voltage source for the load, as shown in Fig. 2. In both cases Z_n is the source or reference impedance. The Thevenin-equivalent model for a load utilizing a dependent voltage source, as shown in Fig. 2, is similar to the standard technique for modeling the input z-parameters of a two-port, where $Z_n^* = Z_{11}$ and $V_o = Z_{12} I_2$.

The maximum power available to Z which corresponds to the power delivered to Z_n^* in Fig. 2b with $V_o = 0$, is

$$|a_n|^2 = \frac{|V_n + Z_n I_n|^2}{4\,|\mathrm{Re}Z_n|} \tag{10}$$

or

$$a_n = \frac{V_n + Z_n I_n}{2\sqrt{|\mathrm{Re}Z_n|}} \quad . \tag{11}$$

The relation for b_n can also be readily found from the Thevenin model of Fig. 2b where V_o is the reflected-voltage generated by any difference of Z from the value of Z_n^*. The

relationships for a_n and b_n are also obtainable by subtracting the power dissipated in Z from that power available from the source.

Kurokawa relates the power waves using a S-parameter matrix, which may be written for a two-port as

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \tag{12}$$

or

$$b = Sa. \tag{13}$$

Setting $Z_n = Z_o$ (pure real), Eq (8) and (9) reduce to

$$a_n = \frac{V_{ni}}{\sqrt{Z_o}} \quad , \quad b_n = \frac{V_{nr}}{\sqrt{Z_o}} \; . \tag{14}$$

Substituting these relations into Eq (13), one obtains

$$\begin{aligned} \frac{V_{1r}}{\sqrt{Z_o}} &= S\,\frac{V_{1i}}{\sqrt{Z_o}} + S\,\frac{V_{2i}}{\sqrt{Z_o}} \\ \frac{V_{2r}}{\sqrt{Z_o}} &= S\,\frac{V_{1i}}{\sqrt{Z_o}} + S\,\frac{V_{2i}}{\sqrt{Z_o}} \quad . \end{aligned} \tag{15}$$

Comparing Eqs (15) and (7), it is obvious that the previously formulated S-parameters of Eq (7) are actually the scattering parameters of Dicke and Kurokawa.

To measure S_{11}, a_2 must be zero such that

$$S_{11} = \left.\frac{b_1}{a_1}\right|_{a_2 = 0} . \tag{16}$$

This is done by placing a signal at port-1 and measuring the ratio of the reflected to incident power waves at port-1, with port-2 terminated in the reference load, Z_o. With a reference load on port-2, all the power traveling down the output transmission line from port-2 to the load is dissipated. Thus, none of the power is reflected which could appear as incident power, $|a_2|^2$ at port-2. With a_2 equal to zero, Eq (16) is a valid measure of S_{11}.

The other S-parameters can be measured using similar methods, all of which involve using only reference loads; not short or open circuit loads. The technical problems involved in manufacturing good reference loads, with constant characteristics over broad frequencies, are measurably less complicated than those required to produce good short or open circuits. This use of reference loads also has the advantage that an active device terminated in a pure real load is less likely to display the type of stability problems that are encountered with the reactive short and open loads. Thus, the probability of instability is greatly reduced.

Physical Significance

A close look at the S-parameters will reveal that each of the parameters directly defines a physical relationship between two of the variables of the two-port shown in Fig. 3. The variables a_n and b_n represent the incident and reflected power-waves at port-n respectively, with a_n and b_n defined

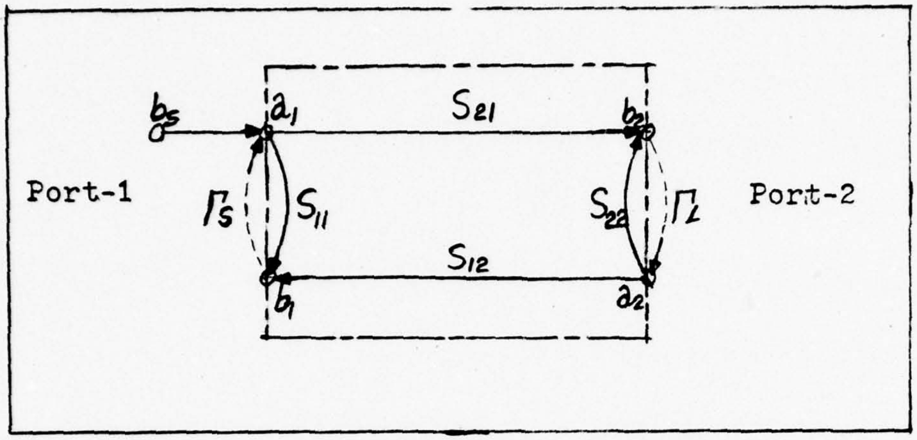

Fig. 3. Signal Flow Diagram of Two Port

in Eqs (8) and (9) as

$$a_n = \frac{V_n + Z_n I_n}{2\sqrt{|ReZ_n|}} , \tag{8}$$

$$b_n = \frac{V_n - Z_n^* I_n}{2\sqrt{|ReZ_n|}} \tag{9}$$

where V_n and I_n represent the total voltage and total current at port-n.

Since there is a linear relationship between a_n, b_n, and V_n, I_n, it can be shown that V_n and I_n are given in terms of a_n and b_n as

$$V_n = \frac{1}{\sqrt{|ReZ_n|}} (Z_n^* a_n + Z_n b_n) \tag{17}$$

and

$$I_n = \frac{1}{\sqrt{|ReZ_n|}} (a_n - b_n) \tag{18}$$

as previously shown the a_ns are related to the b_ns by the

S-matrix of Eq (13)

$$[b] = S[a] \tag{19}$$

or

$$b_1 = S_{11}a_1 + S_{12}a_2 \tag{20}$$

$$b_2 = S_{21}a_1 + S_{22}a_2 \tag{21}$$

which are linear relationships between the power-waves.

As previously shown, the available power (Pa) from a reference source is related to a_1 by

$$P_a = |a_1|^2 \tag{22}$$

which also represents the incident power at port-1 (P_{i_1}) of the two-port in Fig. 3. Similarly the incident power at port-2 is given by

$$P_{i_2} = |a_2|^2 \; . \tag{23}$$

Consequently $|a_n|^2$ is directly related to the incident or available power at port-n.

The power leaving port-n, referred to as departing power, is given as

$$P_{d_n} = |b_n|^2 \tag{24}$$

where $|b_n|^2$ is often referred to as the reflected power of port-n. It is worth noting, that $|b_n|^2$ represents purely reflected power only when all $a_m = 0$ for all $m \neq n$. For the two-port shown in Fig. 3, the departing power at port-1,

with a signal input only at port-1 ($|a_1|^2 \neq 0$), has two reflection components contributing to the total $|b_n|^2$. One component is due to S_{11} and is given by $S_{11}a_1$. The other component is due to any reflection caused by non-zero reflection coefficients at port-2. Even though a signal source is not driving port-2, the existence of a nonzero power-wave reflection coefficient (Γ_L) caused an apparent a_2 given by

$$a_2 = \Gamma_L b_2 \; . \tag{25}$$

Using Eqs (20), (21), and (25) one obtains

$$b_2 = \frac{S_{21}}{(1 - S_{22}\Gamma_L)} a_1 \tag{26}$$

or a P_{d_2} of

$$P_{d_2} = |b_2|^2 = \left| \frac{S_{21}}{(1 - S_{22}\Gamma_L)} \right|^2 |a_1|^2 \; . \tag{27}$$

Substituting Eq (26) into Eq (25) and then Eq (20) yields a P_{d_1} given by

$$P_{d_1} = |b_1|^2 = \left| S_{11} + \frac{S_{12}S_{21}\Gamma_L}{(1 - S_{22}\Gamma_L)} \right|^2 |a_1|^2 \; . \tag{28}$$

If $\Gamma_L = 0$, which makes $a_2 = 0$, Eqs (20) and (21) yield a measure of S_{11} and S_{21} given by

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2 = 0} , \quad S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2 = 0} \tag{29}$$

where a signal source is placed at port-1 ($a_1 \neq 0$) and port-2 is terminated in a reference load ($a_2 = 0$).

If $a_2 \neq 0$, by placing a signal source on port-2, and $a_1 = 0$, by placing a reference load on port-1, then Eqs (20) and (21) yield a measure of S_{12} and S_{22} as

$$S_{12} = \left.\frac{b_1}{a_2}\right|_{a_1 = 0} , \quad S_{22} = \left.\frac{b_2}{a_2}\right|_{a_1 = 0} \tag{30}$$

all of which confirms the previously mentioned measurement technique.

Since a_n and b_n are vector quantities, this method of measuring each of the S_{kl} yields both magnitude and phase information. The quantity S_{11} represents a forward wave reflection coefficient (Γ_1), and S_{21}, a forward wave gain coefficient (A_f), both of which may be plotted on a Polar chart. The same is true of S_{22} and S_{12} except they represent reverse parameters or Γ_2 and A_r respectively.

If equipment to measure a_n and b_n is not readily available, at least a measure of the magnitudes of S_{kl} can be made by measuring the incident and departing powers. The quantity $|S_{11}|^2$ is the forward power-reflection coefficient (R_f) given by

$$R_f = |\Gamma_1|^2 = |S_{11}|^2 = \frac{P_{d_1}}{P_{i_1}} \tag{31}$$

and $|S_{21}|^2$ is the forward power GAIN (G_f)

$$G_f = |S_{21}|^2 = \frac{P_{d_2}}{P_{i_1}} \tag{32}$$

for a reference terminated network. Likewise $|S_{22}|^2$ is the reverse power reflection coefficient (R_r) and $|S_{12}|^2$ is the reverse power GAIN (G_r) given by

$$R_r = |\Gamma_2|^2 = |S_{22}|^2 = \frac{P_{d_2}}{P_{i_2}}, \quad G_r = |S_{12}|^2 = \frac{P_{d_1}}{P_{i_2}}. \tag{33}$$

To accurately measure the characteristics of a network the angle information for each S_{kl} is vital; however, the previous equations do show the relationships of all the powers entering and leaving a reference terminated two-port.

Amplifiers Characteristics

The previous developments in this thesis have defined the S-parameters and shown how they represent the characteristics of a two-port network. It is important for the designer to be able to derive useful amplifier characteristics such as gain and stability from the S-parameters.

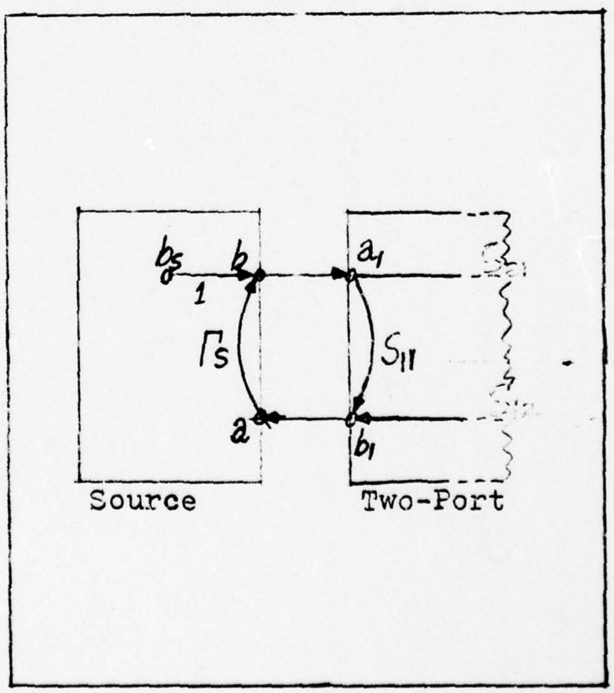

Fig. 4. Signal Source

In deriving an expression for gain, it is advantageous

to reference the variables of the two-port to a standard variable. The maximum power available from a source can provide such a reference variable. A signal source connected to a two-port can be represented as shown in Fig. 4. The power provided is maximum when the power wave reflection coefficient, looking into the input of the two-port is $\Gamma_{in} = \Gamma_s^*$ for a real reference impedance.[+] The power being delivered by the source is

$$P = |b|^2 - |a|^2 \tag{34}$$

where
$$a = \Gamma_{in} b = \Gamma_s^* b \tag{35}$$

and
$$b = b_s + \Gamma_s a \; . \tag{36}$$

Combining these three equations one obtains the maximum power available as (Ref 1:6-3)

$$P = \frac{|b_s|^2}{1 - |\Gamma_s|^2} \tag{37}$$

[+]The power wave reflection coefficient $_{in}$ looking into port n with the input impedance Z_{in} given by V_n/I_n is obtained from Eqs (8) and (9) as

$$\Gamma_{in} = \frac{b_n}{a_n} = \frac{Z_{in} - Z_n^*}{Z_{in} + Z_n} \; .$$

For a termination Γ_L looking out of port n with Z_L equal to $-V_n/I_n$, Γ_L is given by

$$\Gamma_L = \frac{a_n}{b_n} = \frac{Z_L - Z_n}{Z_L + Z_n^*} \; .$$

Gain can be defined in three different ways (Ref 3:6-3), as transducer power gain (G_T), power gain (G), and available power gain (G_A). Transducer power gain is defined as

$$G_T = \frac{\text{Power delivered to load}}{\text{Power available from source}} \tag{38}$$

or

$$G_T = \frac{P_L}{P_a} \, . \tag{39}$$

Referring back to Fig. 3, P_L is the difference in the incident power and the reflected power

$$P_L = |b_2|^2 - |a_2|^2 \tag{40}$$

or

$$P_L = (1 - |\Gamma_L|^2) \, |b_2|^2 \, . \tag{41}$$

Dividing Eqs (41) by (37) gives G_T as

$$G_T = (1 - |\Gamma_s|^2)(1 - |\Gamma_L|^2) \frac{|b_2|^2}{|b_s|^2} \, . \tag{42}$$

From Fig. 3, b_2/b_s can be found to be

$$\frac{b_2}{b_s} = \frac{S_{21}}{(1 - S_{11}\Gamma_s)(1 - S_{22}\Gamma_L) - S_{12}S_{21}\Gamma_s\Gamma_L} \, . \tag{43}$$

Therefore G_T is

$$G_T = \frac{(1 - |\Gamma_s|^2)(1 - |\Gamma_L|^2)|S_{21}|^2}{|(1 - S_{11}\Gamma_s)(1 - S_{22}\Gamma_L) - S_{12}S_{21}\Gamma_s\Gamma_L|^2} \, . \tag{44}$$

Similarly the power gain (G) defined as

$$G = \frac{\text{Power delivered to the load}}{\text{Power input to network}} \tag{45}$$

can be shown to be

$$G = \frac{|S_{21}|^2(1 - |\Gamma_L|^2)}{(1 - |S_{11}|^2) + |\Gamma_L|^2(|S_{22}|^2 - |D|^2) - 2\mathrm{Re}(\Gamma_L[S_{22} - DS_{11}^*])} \tag{46}$$

where $D = S_{11}S_{22} - S_{12}S_{21}$ and likewise the available power gain (G_A) is

$$G_A = \frac{\text{Power available from network}}{\text{Power available from source}} \tag{47}$$

or

$$G_A = \frac{|S_{21}|^2(1 - |\Gamma_s|^2)}{(1 - |S_{22}|^2) + |\Gamma_s|^2(|S_{11}|^2 - |D|^2) - 2\mathrm{Re}(\Gamma_s[S_{11} - DS_{22}^*])} . \tag{48}$$

As a first approach in designing an amplifier one can make the approximation that $S_{12} = 0$. This unilateral approximation allows G_T of Eq (41) to be simplified to give

$$G_{T_u} = \frac{(1 - |\Gamma_s|^2)|S_{21}|^2(1 - |\Gamma_L|^2)}{|1 - S_{11}\Gamma_s|^2|1 - S_{22}\Gamma_L|^2} \tag{49}$$

where G_{T_u} is known as the unilateral transducer power gain. The relationship defined in Eq (49) can be separated into three parts as shown

$$G_{T_u} = \left[\frac{(1 - |\Gamma_s|^2)}{|1 - S_{11}\Gamma_s|^2}\right]\left[|S_{21}|^2\right]\left[\frac{(1 - |\Gamma_L|^2)}{|1 - S_{22}\Gamma_L|^2}\right] . \tag{50}$$

This relation may be written symbolically as

$$G_{T_u} = [G_1] \cdot [G_o] \cdot [G_2] \tag{51}$$

where G_1 represents the gain of the input portion of the two-port, G_o represents the internal gain of an ideal active element, and G_2 represents the gain of the output portion.

When $|S_{11}|$ and $|S_{22}|$ are less than unity, the unilateral transducer gain will be maximum when both input and output ports are conjugately matched, $\Gamma_s = S_{11}^*$ and $\Gamma_L = S_{22}^*$. In this case G_{T_u} becomes

$$G_{T_u}\Big|_{max} = \frac{|S_{21}|^2}{(1 - |S_{11}|^2)(1 - |S_{22}|^2)} . \tag{52}$$

The locus of all Γ_n that provide a specific unilateral power gain (g_n) takes on the form of a circle when plotted on the Smith Chart (Ref 1:3-7). The center of the constant gain circle lays on a line from the center of the chart through S_{nn}^*, where n is the port for which the locus of Γ_n is desired. The distance from the center of the chart to the center of the circle is

$$r_n = \frac{g_n |S_{nn}|}{1 - |S_{nn}|^2 (1 - g_n)} . \tag{53}$$

The radius of the circle is

$$\rho_n = \frac{\sqrt{1 - g_n}\,(1 - |S_{nn}|^2)}{1 - |S_{nn}|^2 (1 - g_n)} \tag{54}$$

where $$g_n = \frac{G_n}{G_n|_{max}} = G_n\,(1 - |S_{nn}|^2)\ . \qquad (55)$$

Therefore specific amplifier gains can be obtained by mismatching the input and output ports as needed.

In using the unilateral approximation, one must have some idea of the magnitude of the error introduced. The unilateral figure of merit (u) as developed by Bodway (Ref 3:6-6) is

$$u = \frac{|S_{11}|\,|S_{22}|\,|S_{12}S_{21}|}{\left|1 - |S_{11}|^2\right|\left|1 - |S_{22}|^2\right|} \qquad (56)$$

and gives a measure sf the error as

$$\frac{1}{|1 + u|^2} < \frac{G_T}{G_{T_u}} < \frac{1}{|1 - u|^2} \qquad (57)$$

where $|\Gamma_s| \leqq |S_{11}|$ and $|\Gamma_L| \leqq |S_{22}|$ for $|S_{11}|$ and $|S_{22}|$ both less than unity. With the limits of G_{T_u} known, it is easy to judge when the error is excessive thus forcing the use of the more complicated G expression. An important point to be made is that G_T is bounded by Eq (57) only for the same matching conditions necessary for G_{T_u} to be maximum. Therefore, $G_T|_{max}$ may occur outside this range since S_{11}^* and S_{22}^* are not necessarily the proper matching conditions for $G_T|_{max}$.

In the unilateral case it was assumed that S_{12} is small and has a negligible effect on G_T. It is of interest to

determine the two-port reflection coefficients when the effects of S_{12} are not negligible (or u not small). The input power wave reflection coefficient can be found from Fig. 3 to be

$$\Gamma_{in} = \frac{b_1}{a_1} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{(1 - S_{22}\Gamma_L)} \tag{58}$$

and similarly the output power wave reflection coefficient is

$$\Gamma_{out} = \frac{b_2}{a_2} = S_{22} + \frac{S_{12}S_{21}\Gamma_s}{(1 - S_{11}\Gamma_s)} . \tag{59}$$

From Eq (58), it is possible that for some $|\Gamma_L|$ less than unity, the $|\Gamma_{in}|$ becomes larger than unity. Likewise from Eq (59) it is possible that for some $|\Gamma_s|$ less than one, $|\Gamma_{out}|$ is larger than unity. Either of these conditions implies that the amplifier can oscillate. For an amplifier to be unconditionally stable $|\Gamma_{in}|$ and $|\Gamma_{out}|$ must be less than unity and as a result a simultaneous match of both input and output ports for maximum power transfer is possible.

The condition of simultaneous matching for maximum power implies that Γ_{in} of Eq (58) and Γ_{out} of Eq (59) are equal to Γ_s^* and Γ_L^* respectively for real reference impedances or

$$\Gamma_s^* = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{(1 - S_{22}\Gamma_L)} \tag{60}$$

and

$$\Gamma_L^* = S_{22} + \frac{S_{12}S_{21}\,\Gamma_s}{(1 - S_{11}\,\Gamma_s)} \quad . \tag{61}$$

Solving Eq (61) for Γ_L and substituting into Eq (54) yields a quadratic equation in Γ_s^*, the solution for which is shown by Anderson (Ref 1:3-12) to be

$$\Gamma_{ms} = C_1^* \left\{ \frac{B_1 \pm \sqrt{B_1^2 - 4|C_1|^2}}{2|C_1|^2} \right\} \tag{62}$$

where

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |D|^2$$

$$C_1 = S_{11} - DS_{22}^* \tag{63}$$

$$D = S_{11}S_{22} - S_{12}S_{21}$$

Solving for Γ_s in Eq (60) and substituting produces a quadratic solution of the form

$$\Gamma_{mL} = C_2^* \left\{ \frac{B_2 \pm \sqrt{B_2^2 - 4|C_2|^2}}{2|C_2|^2} \right\} \tag{64}$$

where

$$B_2 = 1 - |S_{11}|^2 + |S_{22}|^2 - |D|^2 \, . \tag{65}$$

$$C_2 = S_{22} - DS_{11}^*$$

Bodway (Ref 3:6-4) varifies Anderson's equations and shows that if the radical is nonzero then Eqs (62) and (64) each provide two solutions.

The factor under the radical being positive or negative can be associated with a factor $|K|$ being more than or less than unity, where K is defined by Bodway as

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |D|^2}{2|S_{12}|\,|S_{21}|} \,. \qquad (66)$$

If $|K|$ is less than unity, then both $|\Gamma_{ms}|$ and $|\Gamma_{mL}|$ are equal to unity, and represent purely reactive elements. If K is greater than unity, then $|\Gamma_{ms}|$ and $|\Gamma_{mL}|$ are both either less than one are larger than one. If K is less than negative one then either $|\Gamma_{ms}|$ or $|\Gamma_{mL}|$ is less than unity while the other is larger than unity, thus simultaneous matching of input and output is impossible.

The use of K as defined in Eq (66) allows a simplification of the expression for G_A when G_Q is maximum and K is larger than unity, as

$$G_A\Big|_{max} = \left| \frac{S_{21}}{S_{12}} \left(K \pm \sqrt{K^2 - 1}\right) \right| \qquad (67)$$

where (+) is chosen if B_1 of Eq (63) is negative and (-) is chosen if B_1 is positive. The same signs are also used in Eqs (62) and (64).

Froehner (Ref 7:5-2) shows that the boundary of the region of absolute stability is a circle when plotted on the Smith chart. The location of the center of the circle on the chart is

$$r_{sn} = \frac{C_n^*}{|S_{nn}|^2 - |D|^2} \qquad (68)$$

where n is the n^{th} port for which the impedance plane is

being plotted. The radius of the stability circle is

$$\rho_{sn} = \frac{|S_{12}S_{21}|}{|S_{nn}|^2 - |D|^2} \tag{69}$$

For absolute stability this circle must lie completely outside the Smith chart or $|r_{sn}|$ less $|\rho_{sn}|$ must be larger than unity. From this requirement four conditions can be found which when satisfied will assure absolute stability.

$$|S_{11}| < 1 \quad , \quad |S_{22}| < 1$$

$$\left| \frac{|S_{12}S_{21}| - |C_1|}{|S_{11}|^2 - |D|^2} \right| > 1 \tag{70}$$

$$\left| \frac{|S_{12}S_{21}| - |C_1|}{|S_{22}|^2 - |D|^2} \right| > 1$$

It should be apparent that S-parameters can provide the designer with an accurate and powerful tool for designing amplifiers.

III. Basic Design Approach

This chapter provides an explanation of the basic approach to S-parameter design utilizing computer optimization. The parameter independence factor (PIF) is discussed in addition to a method for numerical calculation of the factor. A short discussion of optimization methods and error functions is given both for typical design objectives and for parameter independence. The basic interconnections of two-ports are developed in the context of the associated parameters. Two special configurations for attaching a two-port as a shunt or series one-port element are discussed. The four basic passive components are discussed with their possible two-port configurations and S-parameters given. A review of the philosophy behind the S-parameter design approach is made with emphasis placed on the utility of using S-parameters.

Parameter Independence

Parameter independence is a measure of the sensitivity of network characteristics to variations in the characteristics of an associated device (e.g., transistor or tunnel-diode). A numerical measure of this sensitivity which is called the parameter independence factor (PIF) would be

$$PIF = \frac{2\left|\Delta S_{21}^{A} \big/ S_{21}^{A}\right|^{2}}{\left|\Delta S_{mn}^{D} \big/ S_{mn}^{D}\right|^{2}} \qquad (71)$$

where S_{21}^A is the S_{21} parameter of the amplifier and S_{mn}^D is the S_{mn} parameter of the associated device. If this ratio is taken to the limit about some particular S_{21}^A then PIF becomes

$$PIF_{mn} = \left\{ \left| \frac{\partial S_{21}^A}{\partial Re(S_{mn}^D)} \right|^2 + \left| \frac{\partial S_{21}^A}{\partial Imag(S_{mn}^D)} \right|^2 \right\} \frac{\left| S_{mn}^D \right|^2}{\left| S_{21}^A \right|^2} \qquad (72)$$

where PIF may now be associated with a normalized analytical partial derivative. The ratio $\left| S_{mn}^D \right|^2 / \left| S_{21}^A \right|^2$ of Eq (51) is the normalizing factor for the partial derivative.

For a two-port amplifier, m and n can only take on the values of either 1 or 2, thus the PIF of Eq (72) defines *four possible parameter independence factors*, all of which directly affect the amplifier's primary *function* of providing amplification or gain. Of these four, the factor with respect to S_{21}^A is the most important. A more generalized parameter independence factor than used in this thesis might involve considerations of the partial derivatives of the other amplifier S-parameters with respect to the device S-parameters. Therefore, a completely general PIF would involve sixteen partial derivatives. For the purposes of this thesis, only those four involving the S_{21}^A of the amplifier will be considered since they play a dominate role in the stability of the amplifier's characteristics. For the purposes of analysis, only the partial derivative term with respect to S_{21}^D *of* the device will be used.

Since the parameter independence factor can be

associated mathematically with taking a partial derivative, an analytical solution is possible. However, with the addition of each component to the circuit, the overall S-parameter equations become more complicated, and the original device S-parameters become more enmeshed. Hence, the partial derivative calculations become laborious. As a consequence, to obtain PIF empirically would involve deriving the overall S-parameter, differentiating analytically with respect to the device parameters, and extracting a solution for the particular circuit involved. This procedure does not, in general, lend itself to a simple solution.

One alternative would be to differentiate G_T, Eq (29), to obtain conditions on Γ_S and Γ_L. But one must design input and output matching networks for the corresponding Γ_S and Γ_L, which still neglects any consideration of the trade-off so made with respect to absolute gain. However, by using a finite difference method of obtaining PIF, with the assistance of a digital computer, it is possible to not only do the complex calculations of the overall S-parameters, but the designer is able to impose the necessary constraints to obtain the desired trade-offs in results at either a single or over a broad-band of frequencies.

Each S-parameter is a complex number representing a particular characteristic of a network. In implementing the finite difference calculation of PIF, S_{mn}^D is first shifted by positive and negative percentages only in its magnitude, as shown in Fig. 5, and the difference of the corresponding

percentage shifts of the amplifier S_{21} is compared to these percentage shifts of S^D_{mn}. Secondly, S^D_{mn} is similarly shifted at right angles to the magnitude and a similar normalized difference ratio is calculated. These calculations are similar to partial differentiation with respect to the magnitude and phase of S^D_{mn} rather than the real and imaginary parts as previously mentioned. The sum of the squared magnitudes of these comparisons comprise PIF as given by

$$\mathrm{PIF} = \frac{\left| S^A_{21}\big|_{P2} - S^A_{21}\big|_{P3} \right|^2 + \left| S^A_{21}\big|_{P4} - S^A_{21}\big|_{P5} \right|^2}{(2\Delta_r)^2 \left| S^A_{21}\big|_{P1} \right|^2} \tag{73}$$

where $S^A_{21}\big|_{P1}$ is S^A_{21} corresponding to S^D_{mn} at the point P1 of Fig. 5 and similarly for the other points. The quantity Δ_r is the ratio of the magnitude of the shift in $|S^D_{mn}|$ to $|S^D_{mn}|$ as $|S^D_{mn}|$ is shifted in the four directions. Therefore, with the aid of a digital computer it is feasible for a designer

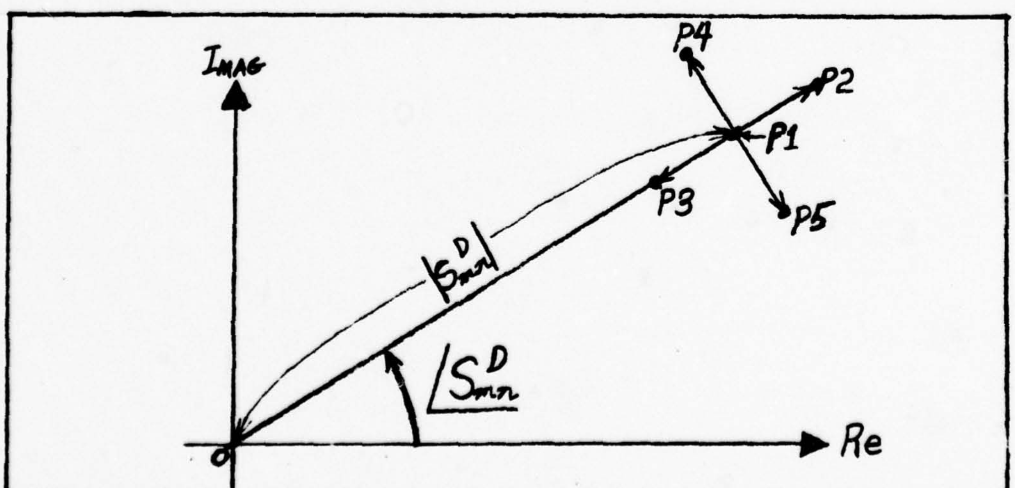

Fig.5 Finite Difference Shifts

to calculate the PIF for any amplifier and include as one of his design criteria an improvement in the amplifier's parameter independence.

Optimization and Error Functions

Any time a designer formulates a circuit design he either builds and tests it, or he models it mathematically. He then judges in what ways its response differs from his requirements, and he makes another circuit or component change in an effort to minimize the error (E) between actual response and desired results. This whole effort to achieve the best or optimum circuit is what is known as optimization, whether the effort is expended by a man or by a computer. In this thesis a direct-search optimization program (PALROS) is used, which is able to rapidly adjust n-components for optimum results based on a user defined error function (E).

If the designer wishes to optimize a circuit for maximum power gain, he could use the error function

$$E = \sum_{freq} \frac{1}{|S_{21}|^2} \tag{74}$$

where $|S_{21}|$ is the magnitude of the overall S_{21} parameter of the network. If flatness or constant gain is desired then

$$E = \sum_{freq} \left| |S_{21}|^2 - 10^{+(G/10)} \right| \tag{75}$$

or

$$E = \sum_{freq} \left| 20 \text{ Log} |S_{21}| - G \right| \tag{76}$$

are possible error function candidates, where G is the desired gain in decibels. For best input and output matching, an E given by

$$E = \sum_{freq} \left[|S_{11}|^2 + |S_{22}|^2 \right] \tag{77}$$

could be chosen to minimize Γ_{in} and Γ_{out}. Trick and Vlach in 1970 (Ref 10:544) used an error function of the form

$$E = \sum_{j=1}^{n} \left[a\left(|S_{11}^{j}|^2 + |S_{22}^{j}|^2 \right) + b\left(|S_{21}^{j}|^2 - |S_{21}^{v}|^2 \right)^d \right] \tag{78}$$

where S_{21}^{v} is a desired gain and d is typically equal to two. The variable exponent d determines the error functions sensitivity to gain deviations.

In this thesis the following error function is used:

$$E = \sum_{NF=1}^{n} \left[\frac{(A+1)}{|S_{21}|^2} + B(PIF) + C \left| |S_{21}|^2 - 10^{+(G/10)} \right| + D\left(|S_{11}|^2 + |S_{22}|^2 \right) + F \left| 10 \text{ Log} |S_{21}|^2 - G \right| \right]. \tag{79}$$

The relative values assigned to A through F determine the weighting or sensitivity of the design to the desired characteristics. The gain may be given versus frequency, thus one can actually define the exact shape of the frequency response curve desired. Since S_{11} and S_{22} represent the input and output power wave reflection coefficients

$$S_{11} = \Gamma_{in} = \frac{Z_{in} - Z^*_{ref}}{Z_{in} + Z_{ref}} \, , \; S_{22} = \Gamma_{out} = \frac{Z_{out} - Z^*_{ref}}{Z_{out} + Z_{ref}} \qquad (80)$$

they will be zero when the input and output are conjugately matched to Z_{ref}. Thus, by specifying D to be nonzero, one includes input and output matching as a measure in the optimization.

<u>Modeling of Two-Port Networks</u>

The characteristics of a two-port network can be modeled by the use of Z, Y, h, g, or S-parameters. For a single two-port, the choice is totally arbitrary, but when two-ports are to be connected in some given manner, the most useful choice of parameters is determined by the configuration in which they are to be connected. If the two-ports are to be connected in series, using Z-parameters will allow the overall Z-parameters of the combination to be determined by simple addition of the respective Z-parameters. The standard two-port circuit configurations are given in Fig. 6. The corresponding parameters associated with each configuration in Fig. 6 require only simple addition to obtain the overall network parameter. The relationships between these four basic parameter sets and S-parameters are developed in Appendix A.

There are three other two-port configurations which are of interest. The first of these involves the cascading of two-ports. Transmission parameters (T) or ABCD parameters are usually used for cascading, both of which lend them-

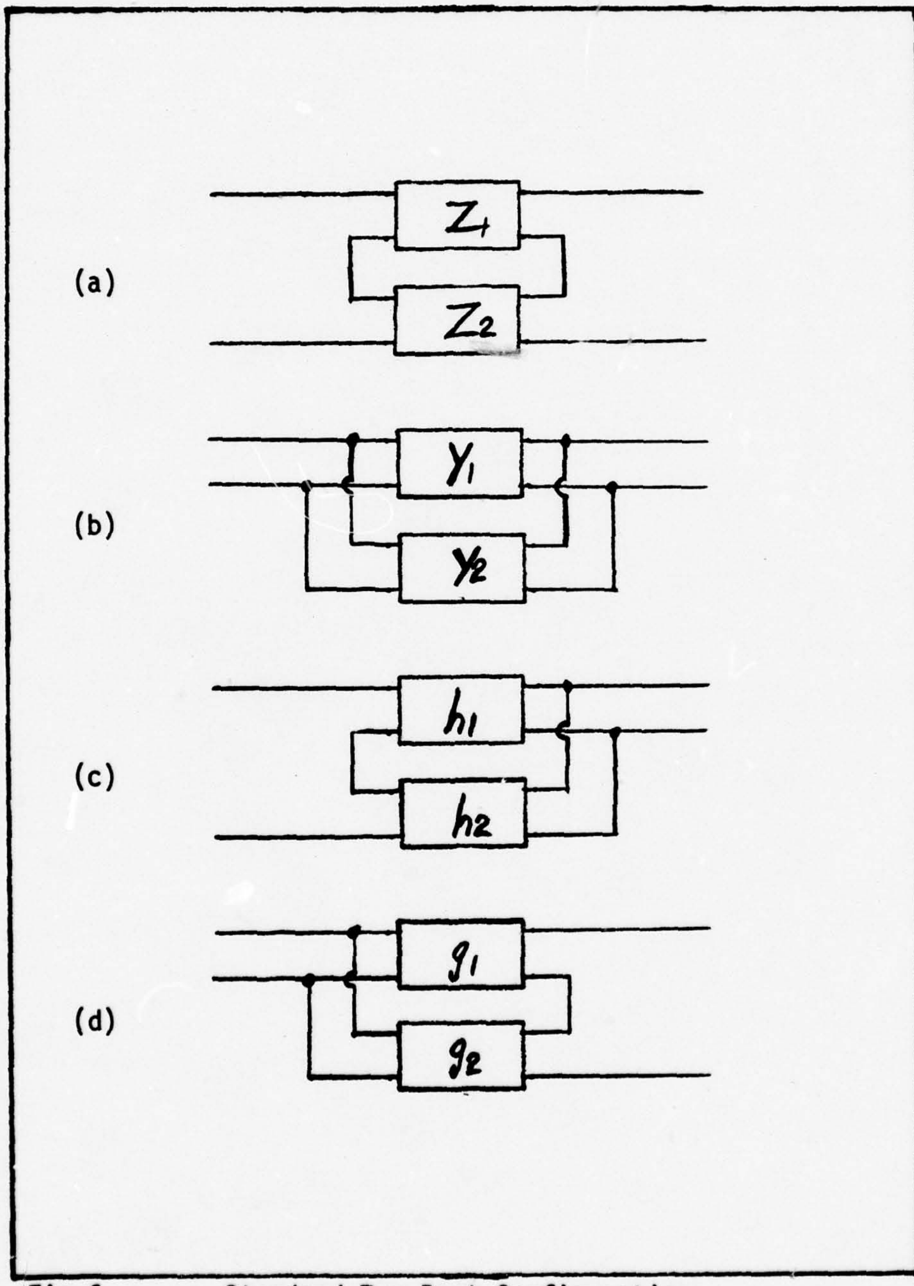

Fig.6 Standard Two-Port Configuration

selves to direct matrix multiplication. One of the characteristics of the numerical procedure used in this thesis is that all intermediate solutions are stored in S-parameter form. Consequently, the S-matrices of the two-ports must both be converted to T or ABCD matrices, the resulting matrices are then multiplied and converted to an overall S-parameter matrix. A direct-mapping technique using only S-parameters allows for more efficient computation of the overall S-matrix for cascaded two-ports as shown in Fig. 7. The direct-mapping technique used in this thesis results in the S-parameters for a real reference impedence given by

$$S^T_{11} = S^1_{11} + \frac{S^1_{12}S^1_{21}S^2_{11}}{1 - S^1_{22}S^2_{11}} , \quad S^T_{12} = \frac{S^1_{12}S^2_{12}}{1 - S^2_{11}S^1_{22}}$$

$$S^T_{21} = \frac{S^1_{21}S^2_{21}}{1 - S^2_{11}S^1_{22}} , \quad S^T_{22} = S^2_{22} + \frac{S^2_{12}S^2_{21}S^1_{22}}{1 - S^2_{11}S^1_{22}} \tag{81}$$

where S^1, S^2, and S^T are two-ports related as shown in Fig. 7.

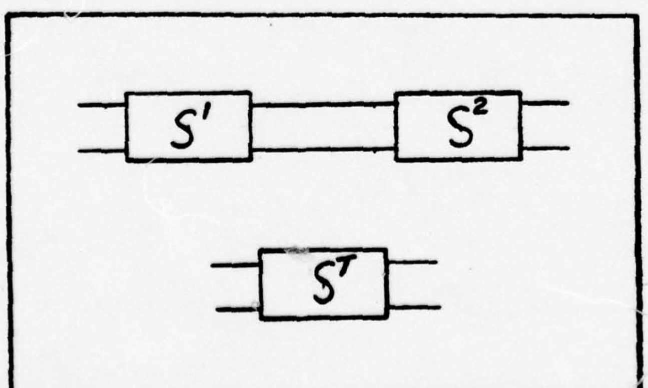

Fig.7 Two-Port Cascade Configuration

The remaining configurations consist of parallel and series connections of two-ports as one-port elements, as shown in Figs. 8 and 9. Two examples of applications of these configurations are series or parallel matching stubs and bias networks.

The parallel two-port circuit shown in Fig. 8 is developed as a one-port element Y'_{in} given as

$$Y'_{in} = Y_{11} - \frac{Y_{12}Y_{21}}{Y_{22}} \tag{82}$$

where the Y_{mn} are the Y-parameters of the original two-port. The overall S-parameters for the new two-port (see dotted line in Fig. 8) are the same as a shunt element of Y'_{in} and given by

$$S = \frac{1}{Y'_{in} + 2}\begin{bmatrix} -Y'_{in} & 2 \\ 2 & -Y'_{in} \end{bmatrix} \tag{83}$$

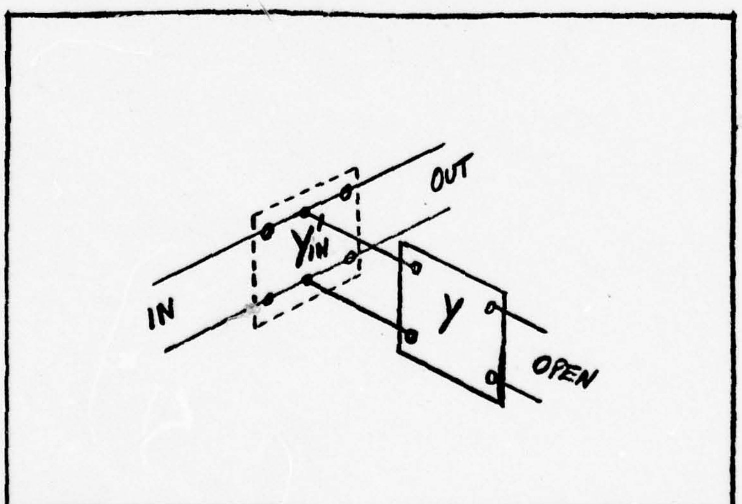

Fig.8 Shunt Two-One-Two Port

where Y'_{in} is normalized to the real reference admittance, $Y_{ref} = \frac{1}{Z_{ref}}$.

The overall S-parameters of the two-port connected in series as shown in Fig. 9 is obtained from

$$Z'_{in} = Z_{11} \tag{84}$$

where Z_{11} is the normalized (1,1) element of the original two-port Z-parameters. The overall S-parameters for the new two-port are the same as that of a series element Z'_{in} and is given by

$$S = \frac{1}{Z'_{in} + 2}\begin{bmatrix} +Z'_{in} & 2 \\ 2 & +Z'_{in} \end{bmatrix}. \tag{85}$$

This technique of converting a two-port to an equivalent one-port and then using the one-port as an element to

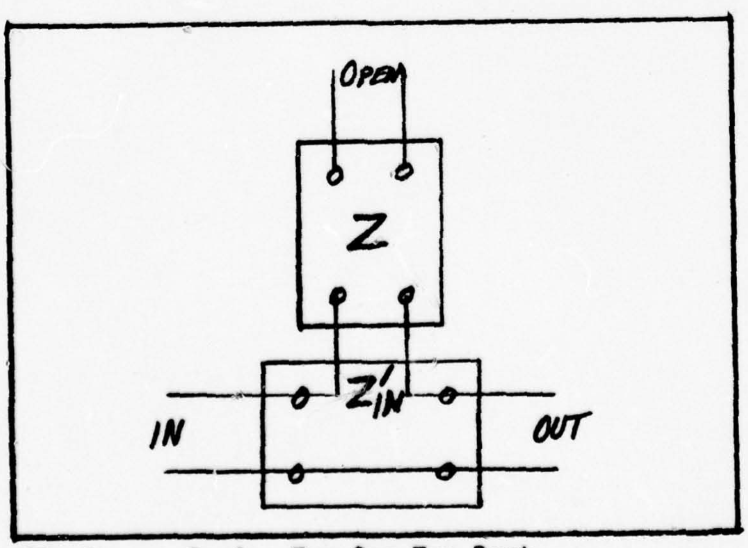

Fig.9 Series Two-One-Two Port

find an equivalent series or parallel element connected as a two-port may also be used to obtain the S-parameters of passive components. Thus one is able to construct any passive two-port network using only four basic types of passive elements or components. These are resistive, capacitive, and inductive lumped components, and the distributed transmission line. The lumped element components can be handled as one-port elements as was the original two-port in Figs. 8 and 9 where Z'_{in} and Y'_{in} are

$$Z'_{in} = \frac{Z_c}{Z_{ref}} = \frac{R_c + jX_c}{Z_{ref}}$$

or

$$Y'_{in} = Y_c Z_{ref} = \frac{Z_{ref}}{R_c + jX_c} \tag{86}$$

where the c subscript designates the component and Z_{ref} is the real reference impedance. Thus, the three basic-discrete components are represented by the S-parameters as defined in Eqs (83) and (85).

The fourth basic component, the transmission line, is itself a two-port network. The S-parameters of a lossless transmission line are defined as

$$S_{11} = S_{22} = \frac{j(Z_o^2 - Z_{ref}^2)\text{Sin}\phi}{2Z_oZ_{ref}\text{Cos}\phi + j(Z_o^2 + Z_{ref}^2)\text{Sin}\phi}$$

and

$$S_{12} = S_{21} = \frac{2Z_oZ_{ref}}{2Z_oZ_{ref}\text{Cos}\phi + j(Z_o^2 + Z_{ref}^2)\text{Sin}\phi} \tag{87}$$

where Z_o is the characteristic impedance of the line, Z_{ref} is the real reference impedance, and ϕ is phase length of the line given by $\phi = wl/v$. The quantities l and v represent the physical characteristics of the line, length and velocity.

Therefore, all four of the basic components can be represented as appropriate two-ports and may be connected in the manners previously discussed to obtain a new overall two-port network. It is now possible using only two-port S-parameter analysis to construct a circuit and find its overall two-port S-parameters. It is worth noting that any component which is not one of the basic components can still be used if the two-port S-parameters can be measured, thus allowing it to be handled as a predefined two-port.

IV. Results

This chapter takes a look at circuits and computer calculations for validation of the computer program, and design method. Several examples are then presented to demonstrate the utility of computer-optimized parameter independent design.

Standard Optimized Design

In order to have some check on computer programing accuracy, to insure the validity of the design method, and to test the coding scheme described in Appendices B through D, three basic validation circuits were utilized. These circuits are design examples in the available literature which use standard techniques to give known responses. The three basic circuits were a 500 megahertz discrete-component amplifier (Ref 6:5-1), a 750 megahertz amplifier utilizing distributed transmission line elements (Ref 6:5-1), and a standard Pi-section filter for matching a 5K ohm source to a 50 ohm load at 3.5 megahertz. A fourth circuit was also used to illustrate a broad-band amplifier design using S-parameter data from the RCA manual (Ref 11:255). In the latter, the original circuit configuration was so complex that a simplified portion of the amplifier was chosen realizing that in doing so the component values listed would not necessarily be optimal. The S-parameters of the devices

used in all the examples are listed in Table I.

Computer calculations were made for each of the four circuits. The 500 megahertz amplifier was originally designed for a flat gain of 12 decibels (db) using the circuit of Fig. 10. Both published and arbitrary component values

Table I
Device S-Parameters

	$f(MH_z)$	S_{11}	S_{21}	S_{12}	S_{22}
2N 3570	500	+ 0.221 -i0.315	+ 0.561 +i2.64	+ 0.000 +i0.045	+ 0.796 -i0.397
	750	+ 0.143 -i0.238	+ 0.842 +i1.73	- 0.004 +i0.078	+ 0.726 -i0.437
2N 3866	150	- 0.1383 -i0.0217	+ 1.1471 +i5.1744	+ 0.0167 +i0.0186	+ 0.1554 - 0.7848
	200	- 0.1675 -i0.0289	+ 1.4754 +i3.3931	+ 0.0295 +i0.0239	+ 0.0913 -i0.7545
	250	- 0.1906 +i0.0881	+ 1.525 +i2.3483	+ 0.0419 +i0.0272	- 0.1265 -i0.6274

were used as initial values in the optimization program. The optimization routine was allowed to search a wide range of possible component values and in both cases the computer calculations confirmed the original values shown in Fig. 10. The largest component differences were for the random data case, where the 7.66 pf capacitor was increased to 34 pf. The largest component difference, for the case where the published values were used as a starting point, was an increase to 8.6 pf for the same 7.66 pf capacitor. All other values for this case were close to their published values.

In both cases the computed gains were 12 db± 0.01 which is in agreement with the original design requirements for this circuit configuration. It is clear that the 7.66 pf capacitor has a negligible effect on the amplifier performance.

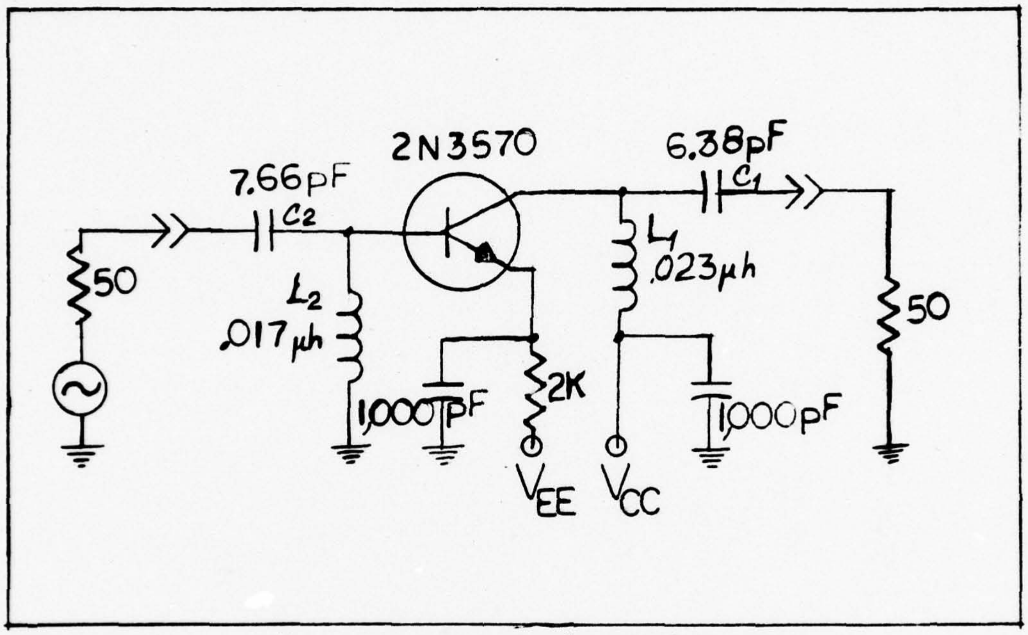

Fig. 10. Discrete Component (500MHz) Amplifier

The 750 megahertz amplifier, shown in Fig. 11 was designed by Froehner (Ref 6:5-1) for maximum power gain, with 12.807 db obtained. The power gain obtained using the computer optimization approach of this thesis was 12.805 db. The largest component change required the lengthening of the output matching stub from 0.715 cm to 0.74 cm. All other line lengths were within 1% of the published values.

The standard Pi-section filter shown in Fig. 12 was designed for no power loss at 3.5 megahertz and 100 db loss at 7.0 megahertz. After optimization for both gain and im-

pedance matching, the program yielded results in reasonable agreement with published values. The largest change in component value occurred for the output capacitor (C_2), a computed value of 203 pf versus an expected value of 875 pf and a loss of 9.4 db at 3.5 megahertz versus a loss of 100.2 db at 7.0 megahertz. As an additional exercise the above

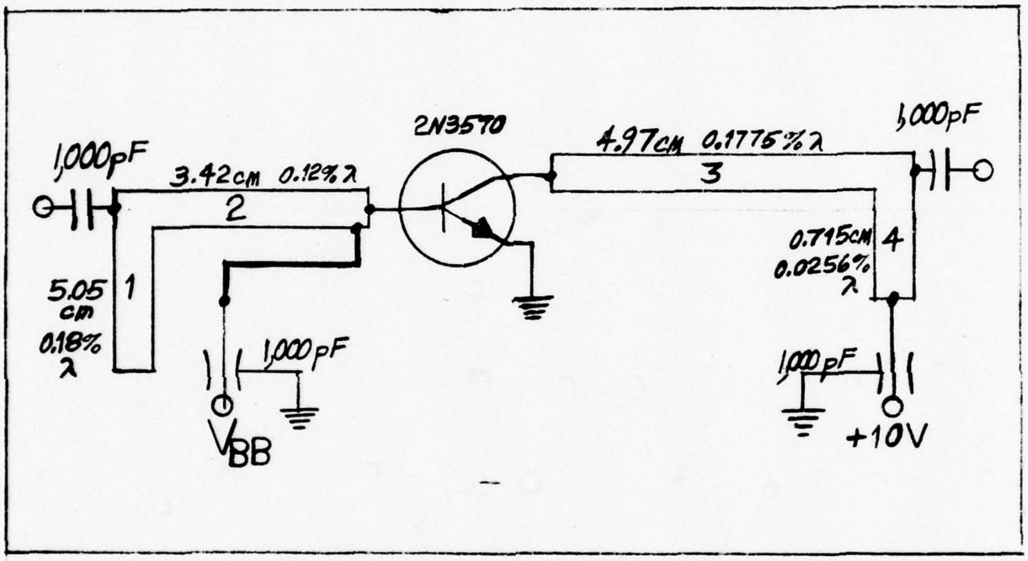

Fig. 11. Distributed Component (750MHz) Amplifier

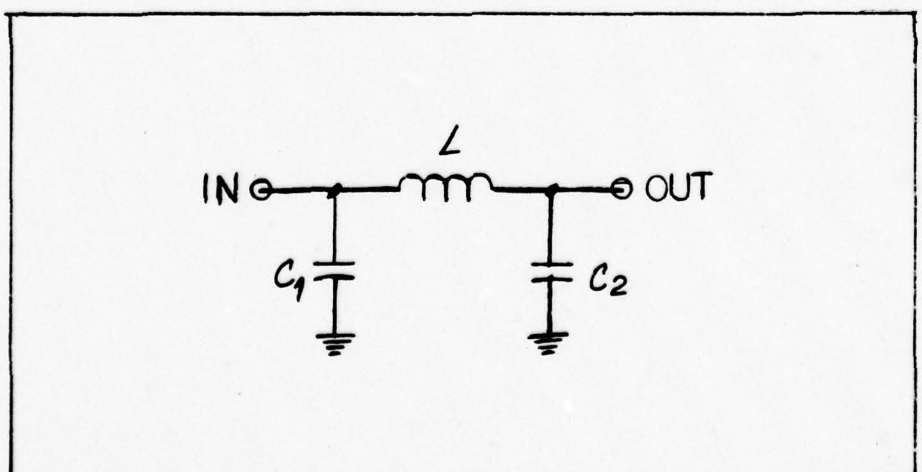

Fig. 12. Pi-Section Filter

calculations were made for a 50 ohm to 50 ohm Pi-section filter which yielded losses of 0.3 db and 100.0 at 3.5 megahertz and 7.0 megahertz respectively. It is interesting that in this case such close conformity to desired losses was possible considering that due to the chosen optimization function and coding scheme, the optimization routine attempted to achieve impedance matching at both frequencies and not just at 3.5 megahertz. The above results are considered sufficient confirmation for this case.

The broad-band amplifier of Fig. 13 (Ref 11:255) using a 2N3866 was optimized for maximum power gain, impedance matching, and then for flatness of gain (specified both in db and as a ratio). The results of the calculations made utilizing various combinations of the above requirements are summarized in Figs. 14 and 15. As shown, when this circuit was optimized for flatness good results were obtained and when optimized for maximum power gain the gain roll-off with increasing frequency was as would be reasonably expected. Several attempts were made to include both flatness and matching as a requirement during optimization. In general, the trade-offs were too severe and a large departure from the desired gain was experienced. These two requirements are incompatible in that to achieve flatness requires selectively mismatching at high gain points and matching at low gain points while "matching" as a requirement dictates equal matching across the band width of the amplifier

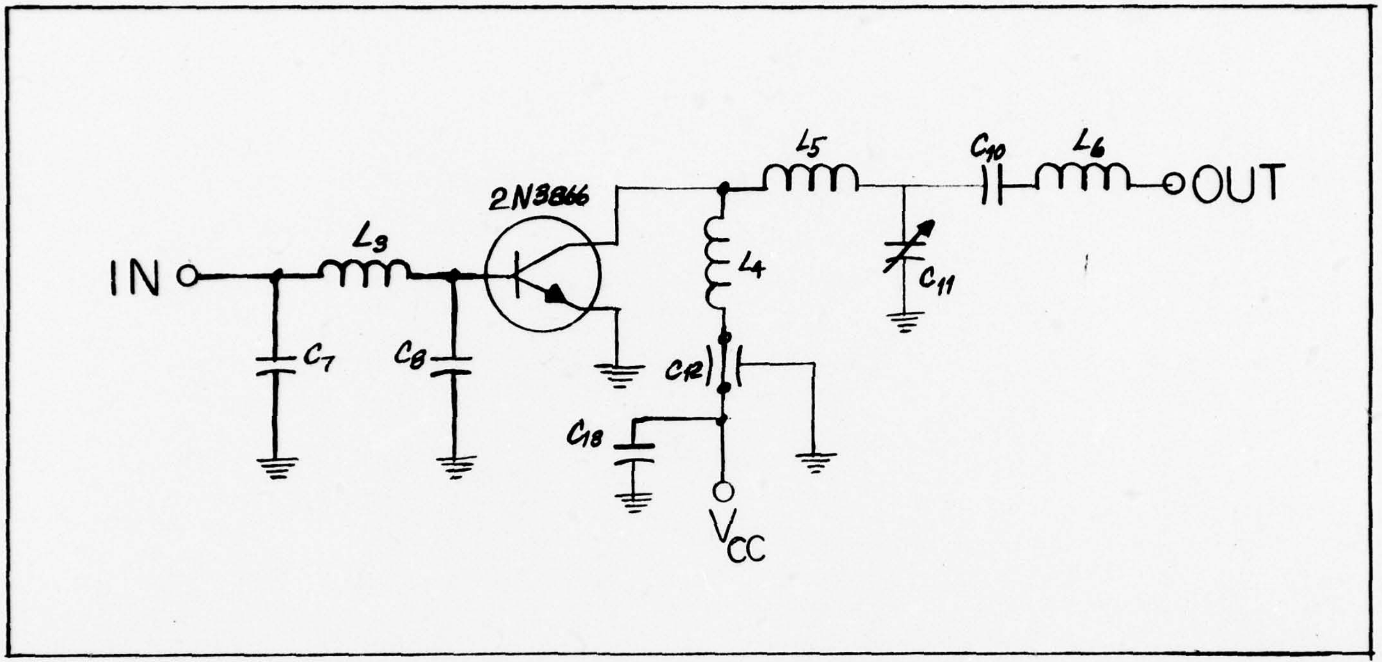

Fig. 13. Broad Band Amplifier

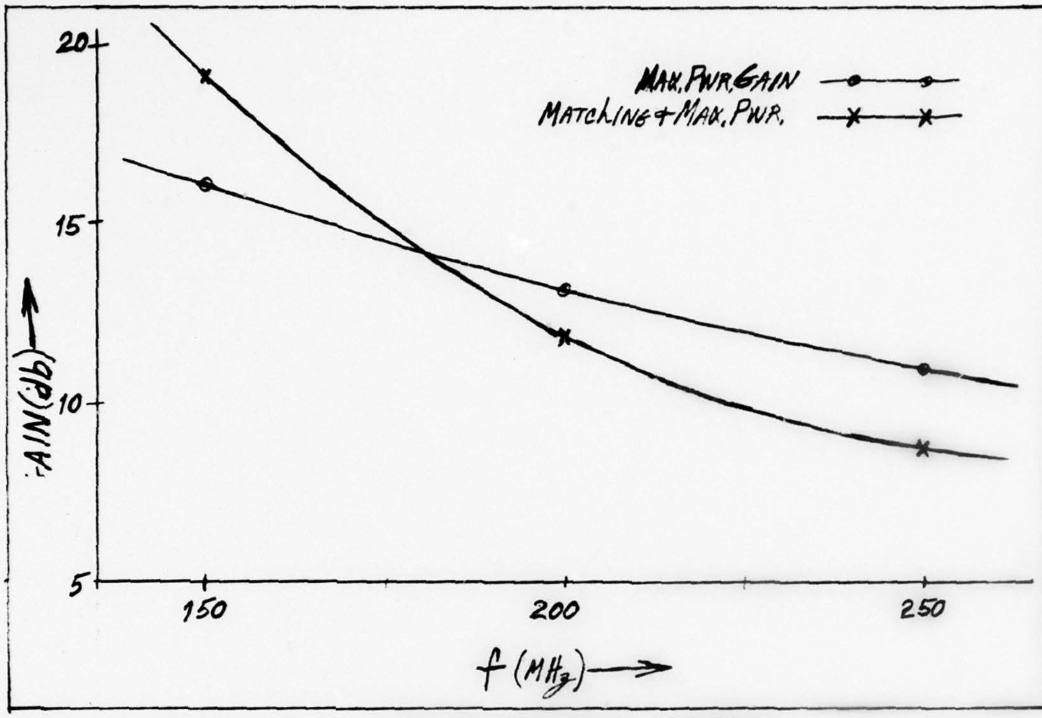

Fig. 14. Maximum Power Gain and Impedance Matching

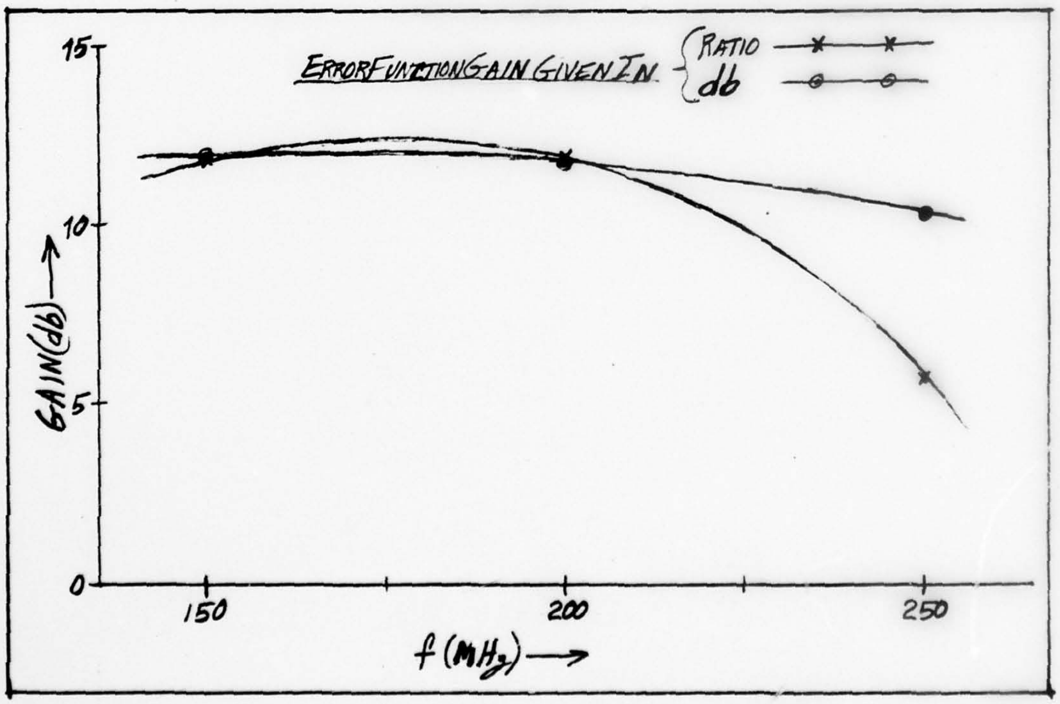

Fig. 15. Flatness (11.8db) As A Ratio Or In DB

(maximum power gain). Therefore, matching should not reasonably be expected for the constant-gain broad-band amplifier.

Parameter Independent Techniques

Although variations in the S-parameters of a nonactive device are possible and probable, only variations with respect to an active device are considered here. Thus, the parameter-independence-factor (PIF) calculations and the optimizations with respect to PIF were made only on the three amplifier circuits.

Calculations of the response of the 500 megahertz discrete-component amplifier optimized with respect to fixed gain (GDB = 12 db) and PIF were made and the results are shown in Table II using random or book data as indicated for program initialization. For a PIF weighting in Eq (79) of

Table II
Discrete Component Amplifier Results

	Random Data	Book Data				
PIF(B)	0.0	0.0	10.0	27.0	27.665	28.0
GAIN(db)	11.999	12.003	11.999	11.810	9.322	6.924
L_1(uh)	0.027	0.023	0.023	0.039	0.042	0.043
C_1(pf)	5.543	6.099	6.106	1.240	1.065	1.000
L_2(uh)	0.012	0.020	0.020	0.030	0.038	0.052
C_2(pf)	33.968	8.578	8.610	99.572	100.0	27.608
IPIF	-	-	2.005	2.005	2.005	2.005
FPIF	-	-	2.005	0.955	0.720	0.594

B = 27.0, the gain was down only 0.2 db while PIF improved from 2.005 to 0.955, over a 2:1 improvement. For B = 27.665 the gain was down 2.68 db while PIF was 0.72 (approximately a 3:1 improvement). In other words, for less than a 3 db loss in gain the parameter independence improved threefold. The amount of loss in gain that can be allowed versus an increase in parameter independence (decrease in PIF) for any circuit is dependent on the circuit specifications, the maximum circuit gain available before considering parameter independence, and the sensitivity of the circuit PIF to component value changes.

Calculations for the distributed 750 megahertz amplifier were optimized with respect to maximum power gain and PIF. As seen in Table III: when the maximum power gain was

Table III
Distributed Component Amplifier Results

Output	Book Data	Random Data			
PIF(B)	0.0	0.0	0.01	0.05	1.0
GAIN(db)	12.805	12.804	12.063	10.883	5.189
L_1(cm)	5.02	8.94	11.41	11.73	9.09
L_2^*(cm)	3.44	14.02	13.52	16.14	18.08
L_3(cm)	4.96	6.38	6.90	7.03	5.04
L_4^*(cm)	0.74	13.29	12.75	12.66	14.84
INITIAL PIF	-	-	8.85	8.85	8.85
FINAL PIF	-	-	2.54	1.73	0.85

*Data contains a half-wavelength anomoly of 14cm.

the only consideration, the gain was 12.8 db and the PIF was 8.85. With the weighting on PIF of only 0.01 the gain dropped only 0.7 db while the PIF had a 3:1 improvement. With a weighting of 0.05 the gain was 10.88 db with an improvement of over 5:1 in PIF. Therefore, for this distributed amplifier dramatic increases in the amplifier parameter independence was achieved with only a minor loss in gain.

The broad-band amplifier was optimized for parameter independence and for both maximum power gain and flatness (11 db) cases with the results plotted in Figs. 16 and 17 respectively. For the maximum power gain case four values of PIF weighting were used. With a PIF weighting of B = 0.0 the gain varied 19.8 db at 150 MHz down to 11.8 db at 250 MHz. For a PIF weighting of B = 1.0 the gain at 150 MHz declined to 7.0 db, only 5% of the original gain or a loss of 95% in gain while at 250 MHz the gain was down to 6.3 db, only 28% of the original gain or a loss of 72% in gain. For this weighting the PIF declined from 6.912 to 4.586 or only a 34% improvement. For a PIF weighting of B = 3.0 the results were degraded even more. However, a weighting of B = 0.01 caused a loss of only 28% at 150 MHz and 1 1/2% at 250 MHz while the PIF improved 12%. Therefore, only the latter weighting offered a break-even trade-off between loss in gain versus PIF improvement. Such a small overall improvement hardly seems worth the effort, but at least the

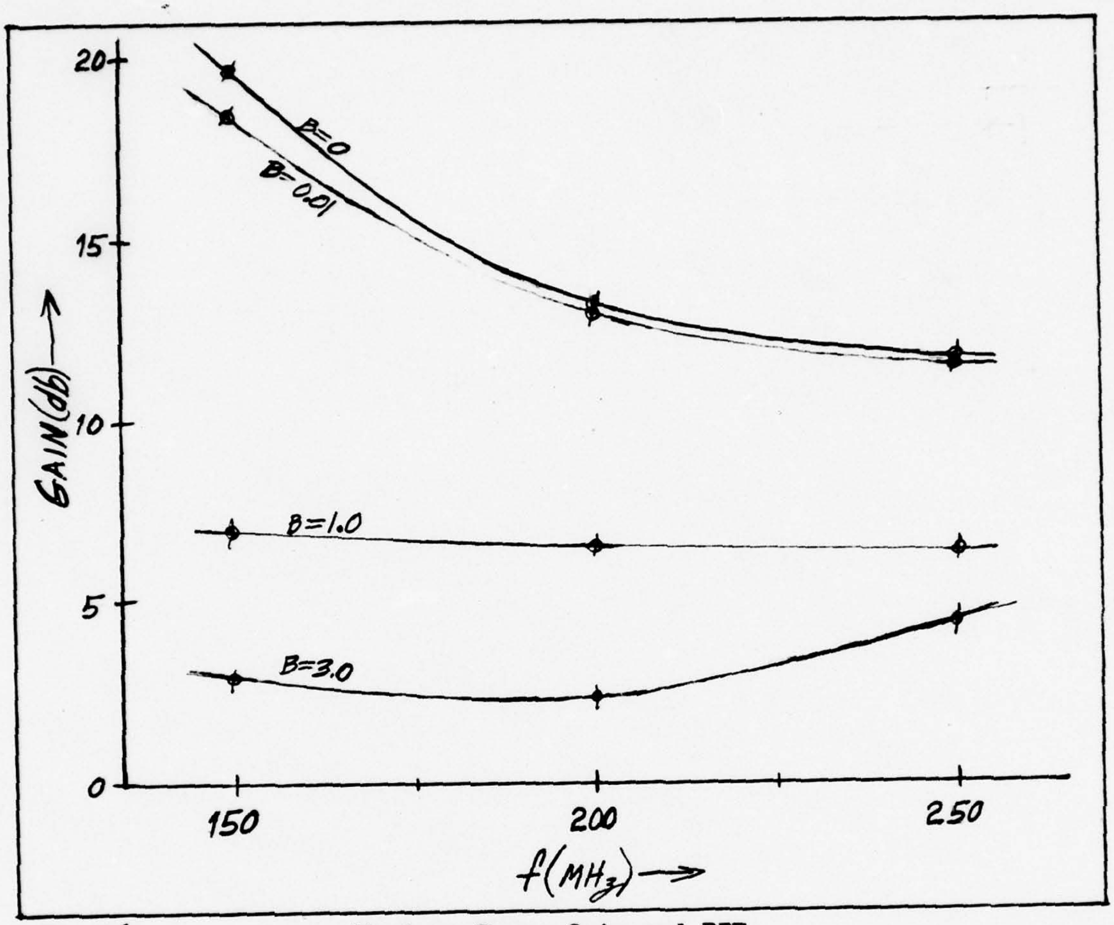

Fig.16. Maximum Power Gain and PIF

PIF for this circuit when optimized for maximum power gain is now known.

For the flat gain case (11 db), placing a weighting of up to 12 on PIF produces less than a 1.4 db loss in gain over the band. A loss of 1.4 db represents 11.0% loss in gain while the PIF only showed an improvement from 5.85 to 3.97 or a 32% improvement. For weightings of 12.3 and above, the loss in gain was over 5 db and the amplifier no

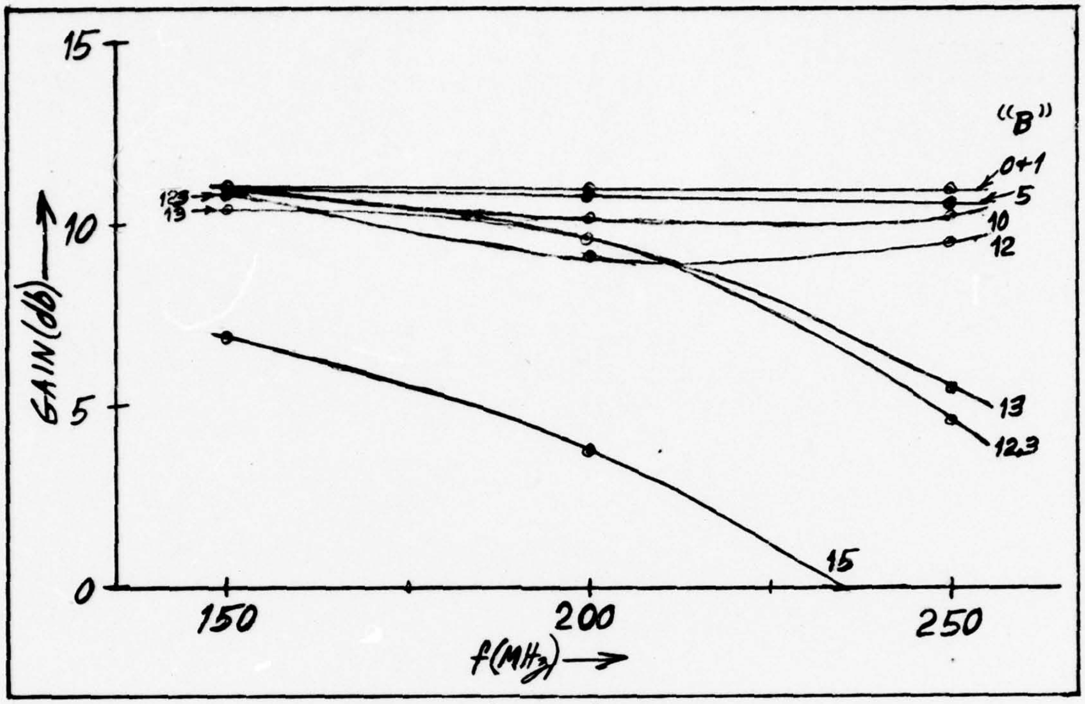

Fig. 17. Flatness (11.0db) and PIF

longer even approximates a constant gain amplifier of 11 db. These results for the broad-band amplifier could well have been expected considering that to achieve flat gain for the broad-band case required many trade-offs in component values such that there were not many component changes possible which would improve the parameter independence without destroying the constant gain capability. The amount of possible improvement in the PIF is limited by the desired gain characteristics. However, as already demonstrated, significant improvement in parameter independence can be obtained with good amplifier characteristics for narrower frequency bands.

V. Conclusions and Recommendations

The feasibility of calculating a parameter independence factor (PIF) for a circuit and the development of a numerical method for such computation on a computer have been demonstrated. Four circuits have been analyzed and the utility of parameter independent design has been demonstrated.

In utilizing the computer program to optimize a design for parameter independence, it was found that the required weighting for parameter independence in the error function that produced optimal results was highly dependent on the particular circuit configuration and frequency band. Single frequency or narrow-band amplifiers required very low weighting while the broad-band amplifier required heavy weighting. When the desired bandwidth of a circuit is increased, the trade-offs required necessarily increase and result in the heaviest weighting on PIF to produce a useful improvement. However, it should be realized that the computer program only evaluates the component values of a circuit that the user specifies; it does not modify the circuit configuration or in general indicate when the circuit should be reconfigured. Even for those circuits which show only negligible improvement in PIF after optimization, the designer knows what the PIF is for that

configuration. He can then evaluate the improvements in PIF that each configuration change makes. It is therefore felt that the concept of including parameter independence as a design consideration is still useful for these cases.

Another feature of the computer aided design portion of this thesis which appears to be unique is the implementation of the conversion between two-ports and one-ports for modeling bias networks and tuning stubs. This conversion method makes it possible to use only four basic passive elements, a resistor, a capacitor, and inductor, and a series transmission line, to model any passive network. A search of available literature reveals that bias networks and tuning stubs are usually handled as special cases requiring up to four additional elements to allow general modeling of passive networks. The use of only four basic elements greatly enhances both the input coding scheme and the ease of utilizing the computer program.

A capability for designing for parameter independence at microwave frequencies utilizing scattering parameters has been demonstrated. The primary impact of this design approach is to make it possible to design microwave amplifiers which will allow for direct field replacement of the active element without requiring extensive realignment.

During the final stages of implementation of the computer program it became obvious that one additional change to the computer program would be of great value to the user

and would enhance the capability of the program to obtain a broader class of solutions. This change requires specifying the error function versus frequency, giving the designer more latitude in specifying the desired characteristics of the network. For example, specifying 0.0 db loss and impedance matching at 3.5 megahertz for a Pi-filter network while requiring only the loss (100 db) at 7.0 MHz with less weighting is an often desired requirement for a transmitter output network.

It is a basic conclusion of this thesis that the design engineer can and should include parameter independence as one of his prime considerations when using S-parameter techniques to design microwave amplifiers.

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Appendix A

Parameter Relationships

As shown in Chapter II, the S-parameters relate the incident (a_n) and reflected (b_n) power waves at a port (n) of a n-port network. For a two-port network, as shown in Fig. 18, the reflected power waves are related to the incident power waves by the S-matrix.

$$b_1 = S_{11}a_1 + S_{12}a_2$$
$$b_2 = S_{21}a_1 + S_{22}a_2 \qquad (88)$$

or

$$b = Sa \qquad (89)$$

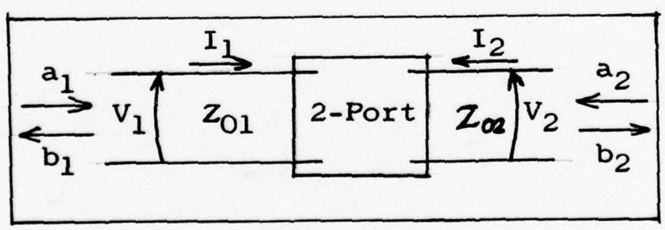

Fig. 18. 2-Port

and the terminal voltages (V_1 & V_2) are related to the terminal currents (I_1 & I_2) by the Z-matrix

$$V_1 = Z_{11}I_1 + Z_{12}I_2$$
$$V_2 = Z_{21}I_1 + Z_{22}I_2 \qquad (90)$$

or

$$V = ZI\ . \qquad (91)$$

In dealing with networks it is advantageous to normalize all measurements and parameters to some reference impedance Z_o. For the multiport network to be considered here, Z_{on}, the reference impedance at port n, is assumed to be real. In normalized form V and I are

$$v_n = \frac{V_n}{\sqrt{Z_{on}}} \quad , \quad i_n = I_n \sqrt{Z_{on}} \tag{92}$$

and Z is

$$z_{ij} = \frac{Z_{ij}}{\sqrt{Z_{oi} Z_{oj}}} \quad .$$

Then in normalized form Eq (91) is given as the normalized Z-parameter matrix (z) must be

$$z = (I + S)(I-S)^{-1} \quad . \tag{96}$$

To find S in terms of z, one simply multiplies on the right by (I-S) and rearranges to obtain

$$(I + z)S = -(I - z)$$

or equivalently

$$S = -(I + z)^{-1}(I - z) \quad . \tag{97}$$

The normalized y matrix is given simply by the inverse of the z matrix

$$y = z^{-1} = (I - S)(I + S)^{-1} \tag{98}$$

and S in terms of y by

$$S = (I + y)^{-1}(I - y) \tag{99}$$

In developing the hybrid case, two new vectors are used (K & L). The vector K is represented by

$$K = \begin{bmatrix} V_1 \\ I_2 \end{bmatrix}$$

and, for the two-port of Fig. 18, vector L is given by

$$L = \begin{bmatrix} I_1 \\ V_2 \end{bmatrix} .$$

These two vectors are related by the H matrix

or
$$K = HL$$

$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} H_{11}, & H_{12} \\ H_{21}, & H_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix} \tag{100}$$

To normalize Eq (100) we desire

or
$$k = hl$$

$$\begin{bmatrix} v_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} h_{11}, & h_{12} \\ h_{21}, & h_{22} \end{bmatrix} \begin{bmatrix} i_1 \\ v_2 \end{bmatrix} \tag{101}$$

and v_n and i_n are given in Eq (92). It is easily shown that the resultant normalized parameters are

$$h_{11} = \frac{H_{11}}{Z_{01}} \quad , \quad h_{12} = \sqrt{\frac{Z_{02}}{Z_{01}}}\ H_{12}$$

$$h_{21} = \sqrt{\frac{Z_{02}}{Z_{01}}}\ H_{21} \ , \ h_{22} = Z_{02}\ H_{22} \ . \tag{102}$$

For the case where $(Z_{01} = Z_{02} = Z_{ref})$ these parameter relationships reduce to

$$h_{11} = \frac{H_{11}}{Z_{ref}} \quad , \quad h_{12} = H_{12}$$

$$h_{21} = H_{21} \quad , \quad h_{22} = Z_{ref} H_{22} \ .$$

Using Eq (94) which related v and i, we obtain

$$k = \begin{bmatrix} v_1 \\ i_2 \end{bmatrix} = \begin{bmatrix} a_1 + b_1 \\ a_2 - b_2 \end{bmatrix} = a - cb$$

and $$l = \begin{bmatrix} i_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} a_1 - b_1 \\ a_2 + b_2 \end{bmatrix} = a + cb$$

where $$c = \begin{bmatrix} -1, & 0 \\ 0, & +1 \end{bmatrix} \ .$$

Substituting for b we have

$$k = (I - CS)a \tag{103}$$

and $$l = (I + CS)a \tag{104}$$

Since k is related to l by h as in Eq (101), we obtain

$$h = (I - CS)(I + CS)^{-1} \tag{105}$$

which is similar to the forms derived for the Z-parameters and Y-parameter cases. To find the S-matrix in terms of h, Eq (105) is solved for S to give

$$S = C(I + h)^{-1}(I - h) . \tag{106}$$

The normalized inverse-hybrid matrix g is given simply by the inverse of the h-matrix

$$g = h^{-1} = (I + CS)(I - CS)^{-1} \tag{107}$$

and S is given by

$$S = -C(I + g)^{-1}(I - g) \tag{108}$$

Appendix B

User's Guide

This appendix provides a user's guide for the reader who wishes to utilize the computer-aided-design program as presented in Appendices C and D.

It was found that writing an interactive program (the computer actively interacting with the user) was too time consuming and such a program would not lend itself to usage on the wide variety of possible user-owned computers. Therefore, as simple a program as possible was sought using the CDC Fortran Extended language. In order to minimize the required computer processing, a data card coding scheme was developed which requires the user to incode his data in a particular format such that the computer knows what each piece of data represents.

This coding scheme defines six basic types of cards (identified by a number 1 to 6 in column one [ID1] of each card) as shown below:

Entry in Column #1	Type of card
1	Used for initialization

Entry in Column #1	Type of card
[Continued]	
2	Used to input the S-parameters of a device
3	Used for defining the values of one of the basic components (resistor, capacitor, inductor, or transmission line)
4 & 5	Used for process control (to tell computer in what way to combine circuit components or circuits)
6	Used to define the error function, used during optimization and as a general end of data card.

The block form of these card types is shown in Table IV. Unformated type read statements are used which allow the user to type in each element separated from the previous element by only a comma. The first two data elements (ID-1 & ID-2) are "Interger" type data and the remaining ten elements are "Real" type data. The following is a card-by-card definition of each abbreviation and an explanation of its function. The "One" cards (ID1=1) come in two kinds, a One-One (ID1=1 & ID2=1) card and a One-Two (ID1=1 & ID2=2) card. The One-One card initializes the following data:

Information Element	Data	Comments
1	1	ID-1 (Type Card)
2	1	ID-2 (Sub-Type)

3	ZREF	Reference Impedence (e.g., 50 ohms)
4	V	Velocity factor for transmission line elements (300 for air line) in mega-meters/sec.
5	ERR	Tolerance in optimized parameters at termination
6	NS	Number of Optimization Request (number of iterations at which to stop if ERR test has not yet been met)
7	KS	Type of Optimization Search 0 - about a "Minimum" 1 - about the first 3 points
8	IMATCH	0 - match input & output to ZREF 1 - match input to Γ_S & output to Γ_L
9	NRR	Number of Random Rotations that the Optimization Program is to perform to insure that the best solution has been found (not just a local minimum of the ERROR FUNCTION)
10	/	End of Data on this card

The One-Two card contains the following information:

Information Element	Data	Comments
1	1	ID-1 (Type Card)
2	2	ID-2 (Sub-Type)
3,4	Γ_S(Real,Imag)	The source reflection coefficient
5,6	Γ_L(Real,Imag)	The load

7	F_n	Frequency at which Γ_s & Γ_L are measured
8	G	Gain desired versus F_n
9	/	End of Data this card

The "2" cards supply the S-parameters for the active devices used. Each device is assigned a module number indicated by element 2. A separate card is specified for each frequency and a separate set of cards for each device. In other words, if an amplifier is to use two transistors and the amplifier's characteristics are to be calculated for three different frequencies, then there should be six (6) type-two cards used.

Information Element	Data	Comments
1 & 2	2, Module #	
3 & 4	S_{11}(Real,Imag)	Device
5 & 6	S_{21}(Real,Imag)	Scattering
7 & 8	S_{12}(Real,Imag)	Parameters
9 & 10	S_{22}(Real,Imag)	
11	F_n	Frequency at which S-parameters measured
12	Para. Tol.	Tolerance of parameters (e.g., = 0.1)
13	/	End of Data this card

There are four types of "three" cards. ID-2 (Information Element 2) identifies the type of component: 1 for a resistor (R), 2 for a capacitor (C), 3 for an inductor (L),

and 4 for a transmission line. Information Element 3 is the initial value of the component and Elements 4 & 5 are the maximum and minimum values respectively that the optimization routines can assign to that component. Element 6, NOPT, designates optimization by the number 2.0 or no optimization by 1.0. Element 7, ISP, designates if the component is to be connected in series or parallel as described in Chapter III. The "three, four" card describes a transmission line with the characteristics impedence Z_o and line length (LL) specified in manner similar to the previous three components. Impedence was allowed to be a variable during optimization to allow for the present practice when working with microstrip transmission lines of varying the strip width.

Type-four and type-five cards are both control cards. The four-card uses ID-2 to identify how two components or circuits are to be combined. The possible combinations for two-port networks were discussed in Chapter III. ID-2 takes on the values 1 for a cascade configuration (Ref Fig. 7), 2 for a series-in/series-out configuration (Z), 3 for a parallel-in/parallel-out (Y), 4 for a series-in/parallel-out (H), 5 for a parallel-in/series-out (G), 6 for a shunt two-one-two port (Ref Fig. 8), and 7 for a series two-one-two port (Ref Fig. 9).

The third and fourth information elements (M1 & M2) are the two modules components or circuits to be combined. As components and modules are read in, the first is

converted to a two-port S-parameter model and is stored in a scratch-pad module (A), while a second is stored in a separate scratch-pad module (B). By placing a code in these element positions (0. for A, -1. for B, or a module number), the user can direct what inputed data or other modules are combined with M1 the left-most element in a cascade configuration. For the two-one-two port configurations, M2 is ignored. After a combination is preformed, the resulting total S-parameter matrix is then stored in A unless A does not appear. In the latter case the result is input to scratch module B. In the two-one-two port conversions, the result replaces the module used.

The five card is used to label sections of a circuit as a module starting with (N+1), where N is the number of devices already labeled using two cards. This labeling ability has proven to be of great advantage in that a circuit can be subdivided (e.g., input matching network, output matching network, bias network, etc.) to allow for easier, faster calculation of the overall S-parameters. If none of the component values have changed in a module then the stored S-parameters of the module are used instead of recalculating them. Also, the combination of major subdivisions or circuits requires this labeling in order to accomplish the combinations previously discussed with the simple use of a four card.

The six card (IDI=6) provides the user the ability to specify the desired response. The user can, by assigning a

numerical value to each of the weighting factors, indicate the importance be placed on each characteristic as shown in Eq (79), repeated here for convenience.

$$E = \sum_{NF=1} \left[\frac{(A + 1)}{|S_{21}|^2} + B(PIF) + C \left| |S_{21}|^2 - 10^{+(G/10)} \right| \right.$$

$$\left. + D \; |S_{11}|^2 + |S_{22}|^2 \; + F \left| 10(\mathrm{Log}_{10} |S_{21}|^2) - G \right| \right] \qquad (79)$$

$$\text{where PIF} = 2 \; \frac{\left| \dfrac{\Delta S_{21}^{A}}{S_{21}^{A}} \right|^2}{\left| \dfrac{\Delta S_{mn}^{D}}{S_{mn}^{D}} \right|^2} \qquad (71)$$

With the details provided in this appendix, the reader should be able to input data to the program as listed in Appendix D in the format of Table IV. This will enable the reader to design parameter independent networks up to and including microwave frequencies.

Table IV
Format For Data Cards

1	1	ZREF	V	ERR	NS	KS	IMATCH	NRR	/		
1	2	$R_e(\Gamma_s)$	$I_m(\Gamma_s)$	$R_e(\Gamma_L)$	$I_m(\Gamma_L)$	F_n	G	/			
2	Mod #	S_{11}		S_{21}		S_{12}		S_{22}		F_n	Tol.
3	1	R	R_{max}	R_{min}	NOPT	ISP	/				
3	2	C	C_{max}	C_{min}	NOPT	ISP	/				
3	3	L	L_{max}	L_{min}	NOPT	ISP	/				
3	4	Z_o	Zo_{max}	Zo_{min}	$NOPT_{Z_o}$	LL	LL_{max}	LL_{min}	$NOPT_{LL}$	/	
4	ID2	M_1	M_2	/							
5	Mod #	M									
6	Mod #	B	NTOL	NPAR	A	C	G(db)	D	F	IG	/

Appendix C

Program Description

The computer aided design program as listed in Appendix D performs the functions of calculating the overall S-parameters of a network, the parameter independence factor, and the optimization error function used in a direct search optimization. The flow diagram given in Fig. 19 begins by reading input data in the format described in Appendix B. This data is used by the design portion of the program to calculate the overall S-parameters and other needed quantities.

DESIGN utilizes the subroutines GENS and NETCO to construct the total S-matrix by respectively calculating the component S-matrices and combining these two-port S-matrices to form the total S-matrix. For parameter independent design, the delta shifts are calculated to obtain the parameter independence factor prior to each optimization step.

PALROS performs the direct search optimization tasks required to minimize the desired error function. This subroutine was provided by Dr. William Davis and is based on the direct search techniques of H. H. Rosenbrock. The results of this optimized network are printed out for user information along with the parameter independence factor.

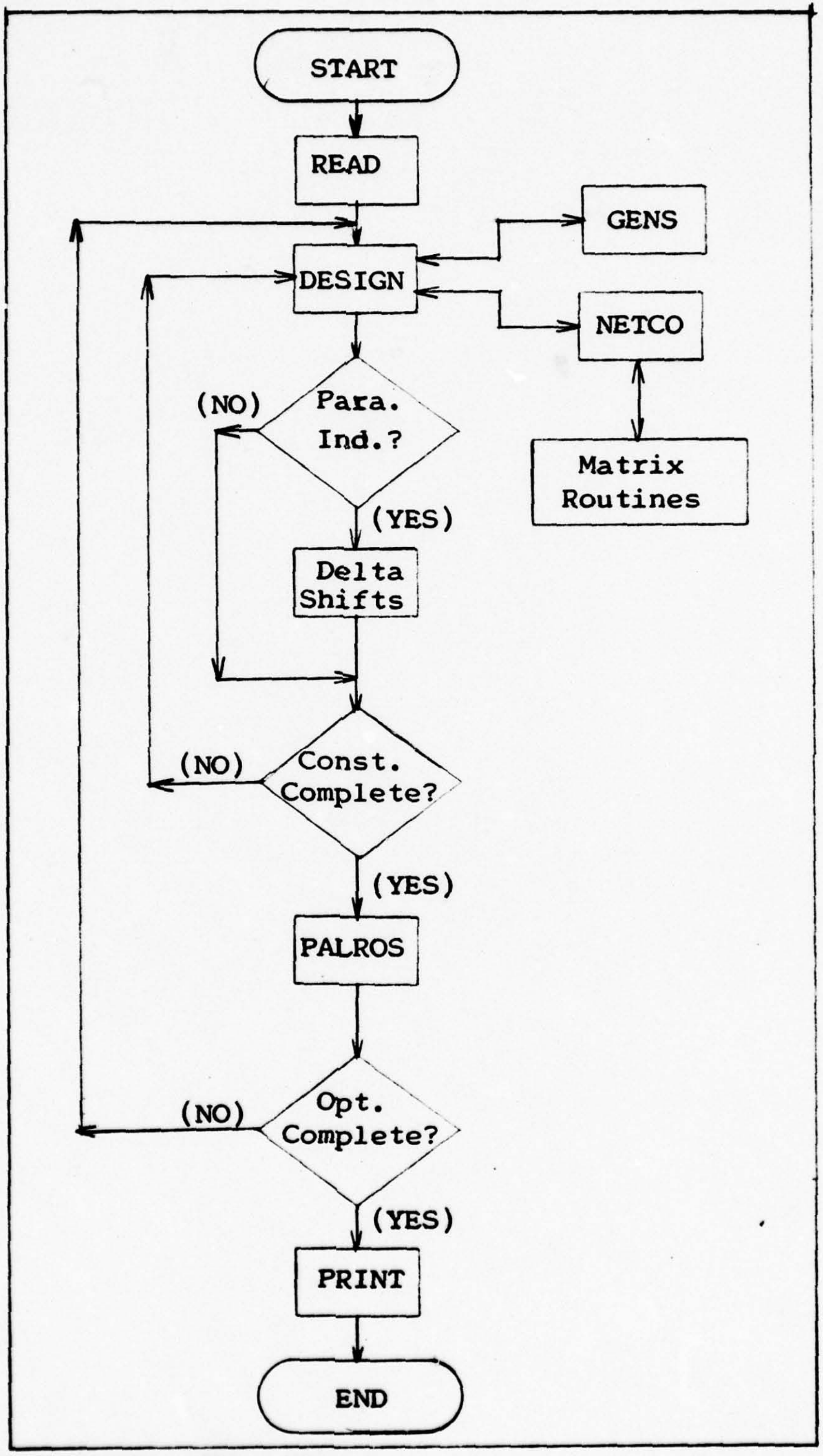

Fig. 19. Flow Diagram

Appendix D

Program Listing

This appendix is a line-by-line listing of the computer program as used on a CDC-6600.

```
      PROGRAM DESIGN(INPUT,OUTPUT,TAPE1=INPUT)

      COMPLEX AMOD(2,2,20),AMPS(2,2,5),CSMA(2,2),
     1CSMB(2,2),D(2,2),GAMMA(10,2),S(4,10,1),G(2,2),F(2,2),
     2GC(2,2),FC(2,2),STOL,C
      DIMENSION CARD(100,4),FREQ(10),TOL(10,1),P(10),QL(10),QU(10),
     1          NP(10,2),IMOD(1),DD(8),GAIN(10)
      EQUIVALENCE(D,DD)
      K=K1=NPO=1 $ M=N=0
    1 DO 100 JD=1,8
  100 DD(JD)=0.
      ID2=0 $ FR=0. $ T=0.
      READ *,ID1,ID2,DD,FR,T
      IF(EOF(1))1001,101
  101 PRINT*,ID1,ID2,DD,FR,T
      GO TO(2,4,6,6,6,15)ID1
    2 IF(ID2.EQ.1) GO TO 3
      M=M+1
      GAMMA(M,1)=D(1,1)
      GAMMA(M,2)=D(2,1)
      FREQ(M)=REAL(D(1,2))
C     FINAL M=# FREQ.S(MAX) & N=FREQ IN USE
      GAIN(M)=DD(6)
      GO TO 1
    3 ZREF=REAL(D(1,1))
      V=AIMAG(D(1,1)) $ ERR=REAL(D(2,1)) $ NS=AIMAG(D(2,1))
      KS=REAL(D(1,2)) $ IMATCH=AIMAG(D(1,2)) $ NRR=REAL(D(2,2))
      GO TO 1
    4 N=N+1 $ TOL(N,K)=T
      IJ=0
      DO 5 J=1,2
        DO 5 I=1,2
      IJ=IJ+1
    5 S(IJ,N,K)=D(I,J)
      IF(M.NE.N)GO TO 1 $ IMOD(K)=ID2 $ K=K+1
C     FINAL K= NO. OF DEVICES
```

```
      N=0 $ GO TO 1
    6 CARD(K1,1)=ID1
      CARD(K1,2)=ID2
      CARD(K1,3)=REAL(D(1,1))
      IF(ID1-4)9,7,7
    7 CARD(K1,4)=AIMAG(D(1,1))
      K1=K1+1 $ GO TO 1
    9 CARD(K1,4)=REAL(D(1,2)) $ K1=K1+1
      IF(ID2.EQ.4) GO TO 11
      IF(AIMAG(D(2,1))-1.5)1,1,10
   10 P(NPO)=ALOG(REAL(D(1,1)))
      OL(NPO)=ALOG(REAL(D(2,1)))
      OU(NPO)=ALOG(AIMAG(D(1,1)))
      NP(NPO,1)=K1-1 $ NP(NPO,2)=3
C     NPO= NO. OF THE CARD NOW IN PROCESSING
C     NP= PARA. IS ON CARD #NPO   , 3 ABOVE=POSITION OF PARA. ON CARD
      NPO=NPO+1 $ GO TO 1
   11 IF(AIMAG(D(2,1))-1.5)13,13,12
   12 P(NPO)=ALOG(REAL(D(1,1)))
      OL(NPO)=ALOG(REAL(D(2,1)))
      OU(NPO)=ALOG(AIMAG(D(1,1)))
      NP(NPO,1)=K1-1 $ NP(NPO,2)=3
      NPO=NPO+1
   13 IF(AIMAG(D(2,2))-1.5)1,1,14
   14 P(NPO)=REAL(D(1,2))
      OL(NPO)=REAL(D(2,2))
      OU(NPO)=AIMAG(D(1,2))
      NP(NPO,1)=K1-1 $ NP(NPO,2)=4
      NPO=NPO+1 $ GO TO 1
   15  NPO=NPO-1 $ IGOTO=-1 $CARD(K1,1)=ID1 $ CARD(K1,2)=ID2
      IF(NPO.LE.0) GO TO 151
      CALL PALROS(P,OL,OU,NPO,NRR,ERR,NS,IGOTO)
  151 K=K-1
      AERR=DD(4)+1. $ BERR=DD(1) $ CERR=DD(5) $ DERR=DD(7) $ FERR=DD(8)
      NTOL=DD(2) $ NPAR=DD(3) $ IG=FR $ GDB=DD(6) $ GS=10.**(GDB/10.)
      IE=0
```

```
C      S-PARA. CALCULATIONS START HERE.
   99 F=0.$ NF=0 $ PIF=0.
  152 IDS=0 $ NF=NF+1 $ ITOL=0
      IA=0
      IF(IMATCH.EQ.0) GO TO 62
      DO 61 I =1,2
      G(I,I)=GAMMA(NF,I)
      F(I,I)=(1-CONJG(G(I,I)))/CABS(1-G(I,I))*SQRT(1-CABS(G(I,I))**2)
      GC(I,I)=CONJG(G(I,I))
   61 FC(I,I)=CONJG(F(I,I))
   62 IF(K.EQ.0)GO TO 160
         DO 16 K2=1,K
      IJ=0
      DO 16 J=1,2
         DO16 I=1,2
      IJ=IJ+1
      IF(IMOD(K2))158,157,159
  157 CSMA(I,J)=S(IJ,NF,K2)
      IA=1
      GO TO 16
  158 CSMB(I,J)=S(IJ,NF,K2)
      IA=2
      GO TO 16
  159 AMOD(I,J,IMOD(K2))=S(IJ,NF,K2)
   16 CONTINUE
  160 CONTINUE
      I=0
  161 I=I+1
      IF(IF.NE.0) GO TO 1000
      ID=CARD(I,1)-2 $ ID1=CARD(I,3) $ ID2=CARD(I,4)
      IF(ID.EQ.2.AND.(NTOL.EQ.ID1.OR.NTOL.EQ.ID2).AND.ITOL.EQ.0)ITOL=I
      GO TO(17,21,41,45)ID
   17 ID=CARD(I,2) $ A=CARD(I,3) $ B=CARD(I,4)
      IF(IA-1)18,19,20
   18 CALL GENS(ID,A,B,CSMA,FREQ(NF),ZREF,V)
      IA=1 $ GO TO 161
```

```
19 CALL GENS(ID,A,B,CSMB,FREQ(NF),ZREF,V)                          000950
   IA=2 $ GO TO 161                                                000960
20 PRINT*,"YOU GOOFED,NEED A CONTROL CARD"                         000970
   GO TO 1000                                                      000980
21 ID=CARD(I,2)                                                    000990
   I1=CARD(I,3) $ I2=CARD(I,4)                                     001000
   IF(ID.LT.6) GO TO 25                                            001010
   IF(I1)23,22,24                                                  001020
22 CALL NETCO(ID,CSMA,CSMB,1,CSMA,IE)                              001030
   IA=1
    GO TO 161                                                      001040
23 CALL NETCO(ID,CSMB,CSMA,1,CSMB,IE)                              001050
   GO TO 161                                                       001060
24 CALL NETCO(ID,AMOD(1,1,I1),    CSMA    ,1,AMOD(1,1,I1),IE)      001070
   GO TO 161                                                       001080
25 IF(I1.EQ. 0.OR.I2.EQ. 0) GO TO 28                               001090
   IF(I1.EQ.-1)GO TO 27                                            001100
   IF(I2.EQ.-1)GO TO 26                                            001110
   CALL NETCO(ID,AMOD(1,1,I1),AMOD(1,1,I2),1,    CSMA    ,IE)      001120
   IA=1
    GO TO 161
26 CALL NETCO(ID,AMOD(1,1,I1),    CSMB    ,1,    CSMB    ,IE)      001130
   GO TO 161                                                       001140
27 CALL NETCO(ID,    CSMB    ,AMOD(1,1,I2),1,    CSMB    ,IE)      001150
   GO TO 161                                                       001160
28 IA=1
   IF(I1.EQ.-1.OR.I2.EQ.-1) GO TO 30
   IF(I1.EQ. 0) GO TO 29                                           001180
   CALL NETCO(ID,AMOD(1,1,I1),    CSMA    ,1,    CSMA    ,IE)      001190
   GO TO 161                                                       001200
29 CALL NETCO(ID,    CSMA    ,AMOD(1,1,I2),1,    CSMA    ,IE)      001210
   GO TO 161                                                       001220
30 IF(I1.EQ. 0) GO TO 31                                           001230
   CALL NETCO(ID,    CSMB,    CSMA,1,    CSMA,IE)                  001240
   GO TO 161                                                       001250
31 CALL NETCO(ID,    CSMA,    CSMB,1,    CSMA,IE)                  001260
```

```
      GO TO 161
   41 IF(CARD(I,3))412,410,414
  410 DO 42 I1=1,2
          DO 42 I2=1,2
   42 AMOD(I1,I2,CARD(I,2))=CSMA(I1,I2)
      IA=0 $ GO TO 161
  412 DO 413 I1=1,2
          DO 413 I2=1,2
  413 AMOD(I1,I2,CARD(I,2))=CSMB(I1,I2)
      IA=1 $ GO TO 161
  414 DO 415 I1=1,2
          DO 415 I2=1,2
  415 AMOD(I1,I2,CARD(I,2))=AMOD(I1,I2,CARD(I,3))
   43 CONTINUE
      GO TO 161
   45 IDS=IDS+1 $ IF(IMATCH.EQ.0) GO TO 46
      CALL CMM(G,CSMA,CSMB)
      CALLCDMA(CSMB,CSMB,-1,+1,+1)
      CALL CMI(CSMB,CSMB,IE)
      CALL CMM(CSMB,FC,CSMB)
      CALL CMAS(CSMA,GC,-1,CSMA)
      CALL CMM(CSMA,CSMB,CSMA)
      CALL CMI(F,CSMB,IE)
      CALL CMM(CSMB,CSMA,CSMA)
   46 IF(CARD(I,2))462,461,464
  461 DO 47 I1=1,2
          DO 47 I2=1,2
   47 AMPS(I1,I2,IDS)=CSMA(I1,I2)
      GO TO 48
  462 DO 463 I1=1,2
          DO 463 I2=1,2
  463 AMPS(I1,I2,IDS)=CSMB(I1,I2)
      GO TO 48
  464 DO 465 I1=1,2
          DO 465 I2=1,2
  465 AMPS(I1,I2,IDS)=AMOD(I1,I2,CARD(I,2))
```

```
   48 CONTINUE
      IA=0
C                          DEL-S SHIFTS
      IF(NTOL.EQ.0.OR.BERR.EQ.0.)GO TO 486
      GO TO (481,482,483,484,485)IDS
  481 C=(0.,1.)
      STOL=S(NPAR,NF,NTOL)*TOL(NF,NTOL)
      S(NPAR,NF,NTOL)=S(NPAR,NF,NTOL)+STOL
      I=ITOL-1 $ GO TO 161
  482 S(NPAR,NF,NTOL)=S(NPAR,NF,NTOL)-2.*STOL
      I=ITOL-1 $ GO TO 161
  483 S(NPAR,NF,NTOL)=S(NPAR,NF,NTOL)+STOL*(1.+C)
      I=ITOL-1 $ GO TO 161
  484 S(NPAR,NF,NTOL)=S(NPAR,NF,NTOL)-2.*STOL*C
      I=ITOL-1 $ GO TO 161
  485 S(NPAR,NF,NTOL)=S(NPAR,NF,NTOL)+STOL*C
  486 P1=1./CABS(AMPS(2,1,1))**2
      IF(BERR.EQ.0..OR.TOL(NF,NTOL).EQ.0.)GO TO 487
      PIF=PIF+(CABS(AMPS(2,1,2)-AMPS(2,1,3))**2+CABS(AMPS(2,1,4)
     1        -AMPS(2,1,5))**2)*P1/(2.*TOL(NT,NTOL))**2
  487 IF(IG.EQ.0)GO TO 488
      GDB=GAIN(NF) $ GS=10.**(GDB/10.)
  488 P2=CABS(AMPS(2,1,1))**2-GS
      P3=20.*ALOG10(CABS(AMPS(2,1,1)))-GDB
       E=E+PIF*BERR+P1*AERR+CERR*ABS(P2)+DERR*(CABS(AMPS(1,1,1))**2+CABS
     1  (AMPS (2,2,1))**2)+FERR*ABS(P3)
                                  .OR.NRR.LE.1)
      IF(IGOTO.GE.4.OR.NPO.EQ.0) PRINT*,((AMPS(I,J,1),I=1,2),J=1,2),E,
     1    PIF,FREQ(NF)
      IF(NF.NE.M) GO TO 152
      IF(NPO.EQ.0) GO TO 900
      IF(IGOTO.GE.4) GO TO 901
      CALL PALROS(P,CL,QU,NPO,NRR,E,KS,IGOTO)
```

```
      DO 53 IP=1,NPO                                                    001520
      IC=NP(IP,1)                                                       001530
      I2=NP(IP,2)                                                       001540
      IF(I2.EQ.4) GO TO 52                                              001560
   51 CARD(IC,I2)=EXP(P(IP))                                            001570
      GO TO 53                                                          001580
   52 IF(I2.EQ.3) GO TO 51                                              001590
      CARD(IC,I2)=P(IP)                                                 001600
   53 CONTINUE                                                          001610
      GO TO 99
  901 IF(IGOTO.EQ.5) PRINT*,"NO. OF FUNCTION-REQUESTS EXCEEDED"         001630
      PRINT*,(P(I),I=1,NPO),E,PIF                                       001640
  900 DO 910 I=1,K1
  910 PRINT*, (CARD(I,J),J=1,4)
 1000 CONTINUE                                                          001650
      GO TO 1
 1001 CONTINUE
      END                                                               001660
```

```
      SUBROUTINE NETCO(ID2,A,B,N,CSM,IE)
      COMPLEX A(2,2),B(2,2),CSM(2,2),S1(2,2),S2(2,2),S3(2,2),P,C
      GO TO (100,200,200,200,200,300,400)ID2
  100 CALL CSMM(A,B,CSM)
      RETURN
  200 CALL CPCFS(S1,A,ID2,IE)
      IF(IE.EQ.0)GO TO 210
      PRINT*,"ERROR",N,ID2,1,"A"
      RETURN
  210 CALL CPCFS(S2,B,ID2,IE)
      IF(IE.EQ.0)GO TO 220
      PRINT*,"ERROR",N,ID2,1,"B"
      RETURN
  220 CALL CMAS(S1,S2,+1,S3)
      CALL CPCTS(CSM,S3,ID2,IE)
      IF(IE.EQ.0)GO TO 230
      PRINT*,"ERROR",N,ID2,2
  230 RETURN
  300 Q=-1. $ GO TO 900
  400 Q=+1.
  900 C=A(1,1)*A(2,2)-A(2,1)*A(1,2)
      P=1-A(2,2)+Q*(C-A(1,1))
      IF(P.EQ.0.) GO TO 1000
      P=(1-A(2,2)-Q*(C-A(1,1)))/P
      C=1./(P+2.)
      CSM(1,1)=CSM(2,2)=C*Q*P
      CSM(2,1)=CSM(1,2)=C*2.
      RETURN
 1000 CSM(1,1)=CSM(2,2)=Q
      CSM(2,1)=CSM(1,2)=0.
      RETURN
      END
```

```
      SUBROUTINE GENS(ID2,D,E,S,FREQ,ZREF,V)
      COMPLEX S(2,2),P,A,B,C
      PI=3.141592654
      P0=ZREF
      GO TO(200,100,100,700,500)ID2
  100 X=2.*PI*FREQ*D
      IF(X.EQ.0..AND.ID2.EQ.2) GO TO 610
      IF(ID2.EQ.2)X=-1./X
C                   THE (-)SIGN IS NECESSARY (1/J=-J).
      P=CMPLX(0.,X)
      GO TO 300
  200 P=CMPLX(D,0.)
  300 IF(E.EQ.2.)GO TO 400
      IF(P.EQ.0.)GO TO 600
      P=1./P
      P0=1./P0
  400 Q=1.
      IF(E.EQ.1.)Q=-1.
      P=P/P0
      C=1./(P+2.)
      S(1,1)=S(2,2)=C*Q*P
      S(2,1)=S(1,2)=C*2.
  500 RETURN
  600 S(1,1)=S(2,2)=-1.
      S(1,2)=S(2,1)=0.
      RETURN
  610 Q=1.
      IF(E.EQ.1.)Q=-1.
      S(1,1)=S(2,2)=(Q+1.)/2
      S(1,2)=S(2,1)=(1.-Q)/2
      RETURN
```

```
700 Z0=0
    BL=F*2.*PI*FREQ/V
    A=CMPLX(0.,(Z0*Z0-ZREF*ZREF)*SIN(BL))
    B=CMPLX(2.*Z0*ZREF*COS(BL),(Z0*Z0+ZREF*ZREF)*SIN(BL))
    S(1,1)=S(2,2)=A/B
    C=CMPLX(COS(BL),(Z0/ZREF)*SIN(BL))
    S(1,2)=S(2,1)=C/(1.+S(1,1))
    RETURN
    END
```

```
      SUBROUTINE CPCFS(A,B,ID2,IE)
      COMPLEX A(2,2),B(2,2),S1(2,2),S2(2,2),S3(2,2)
      GO TO (17,10,12,13,15) ID2
      ENTRY CPCTS
      GO TO (17,11,12,14,16) ID2
   10 CALL CDMA(S1,B,1.,1.,1.)
      CALL CDMA(S2,B,-1.,1.,1.)
      CALL CMI(S2,S3,IE)
      CALL CMM(S1,S3,A)
      RETURN
   11 CALL CDMA(S1,B,1.,-1.,-1.)
      CALL CDMA(S2,B, 1., 1., 1.)
      CALL CMI(S2,S3,IE)
      CALL CMM(S1,S3,A)
      RETURN
   12 CALL CDMA(S1,B,-1., 1., 1.)
      CALL CDMA(S2,B, 1., 1., 1.)
      CALL CMI(S2,S3,IE)
      CALL CMM(S1,S3,A)
      RETURN
   13 CALL CDMA(S1,B,-1.,-1., 1.)
      CALL CDMA(S2,B, 1.,-1., 1.)
      CALL CMI(S2,S3,IE)
      CALL CMM(S3,S1,A)
      RETURN
   14 CALL CDMA(S1,B,-1., 1., 1.)
      CALL CDMA(S2,B, 1., 1., 1.)
      CALL CMI(S2,S3,IE)
      CALL CMM(S1,S3,A)
      A(1,1)=-1.*A(1,1)
      A(1,2)=-1.*A(1,2)
      RETURN
   15 CALL CDMA(S1,B,-1.,-1., 1.)
      CALL CDMA(S2,B, 1.,-1., 1.)
```

```
      CALL CMI(S1,S3,IE)
      CALL CMM(S3,S2,A)
      RETURN
   16 CALL CDMA(S1,B, 1.,-1.,-1.)
      CALL CDMA(S2,B, 1., 1., 1.)
      CALL CMI(S2,S3,IE)
      CALL CMM(S1,S3,A)
      A(1,1)=-1.*A(1,1)
      A(1,2)=-1.*A(1,2)
   17 RETURN
      END
```

```
      SUBROUTINE CMM(A,B,C)
C     THIS S.P. COMPUTES THE PRODUCT(C) OF TWO COMPLEX MATRIXS A&B.
      COMPLEX A(2,2),B(2,2),C(2,2),D,E,F
      D=A(1,1)*B(1,1)+A(1,2)*B(2,1)
      E=A(1,1)*B(1,2)+A(1,2)*B(2,2)
      F=A(2,1)*B(1,1)+A(2,2)*B(2,1)
      C(2,2)=A(2,1)*B(1,2)+A(2,2)*B(2,2)
      C(1,1)=D
      C(1,2)=E
      C(2,1)=F
      RETURN
      END
```

```
      SUBROUTINE CSMM(A,B,CSM)
C     THIS S.P. TAKES 2 COMPLEX(2X2) S-PARA. MATRIXS & COMPUTES A TOTAL
C     S-MATRIX FOR THE TWO IN CASCADE. A IS THE 1ST AND B IS THE 2ND
C     ELEMENT & CSM(COMPOSIT S-MATRIX) IS RETURNED.
      COMPLEX A(2,2),B(2,2),CSM(2,2),C
      C=1.-A(2,2)*B(1,1)
      CSM(1,1)=A(1,1)+(A(1,2)*A(2,1)*B(1,1))/C
      CSM(2,2)=B(2,2)+(B(1,2)*B(2,1)*A(2,2))/C
      CSM(2,1)=(A(2,1)*B(2,1))/C
      CSM(1,2)=(A(1,2)*B(1,2))/C
      RETURN
      END
```

```
      SUBROUTINE CDMA(A,B,U,V,W)
      COMPLEX A(2,2),B(2,2)
      DO 100 I=1,2
         DO 100 J=1,2
100         A(I,J)=B(I,J)*U
      A(1,1)=A(1,1)+V
      A(2,2)=A(2,2)+W
      RETURN
      END
```

AD-A045 460 AIR FORCE INST OF TECH WRIGHT-PATTERSON AFB OHIO SCH--ETC F/G 9/5
PARAMETER INDEPENDENT DESIGN UTILIZING SCATTERING PARAMETERS.(U)
JUN 77 W F DUKE

UNCLASSIFIED AFIT-GE/EE/77-5 NL

2 OF 2
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```
      SUBROUTINE CMAS(A,B,M,D)
C     THIS S.P.COMPUTES THE SUM OR DIFF (D) OF TWO MATRIXS A&B,
C     WHERE THE MODE IS (M=+1) FOR SUM & (M=-1) FOR DIFF.
      COMPLEX A(2,2),B(2,2),D(2,2)
      DO 1 J=1,2
         DO 1 I=1,2
    1       D(I,J)=A(I,J)+M*B(I,J)
      RETURN
      END
```

```
      SUBROUTINE CMI(A,B,IE)
C     THIS S.P.COMPUTES THE INVERSE(B) OF A COMPLEX MATRIX(A),BOTH(2X2)S
      COMPLEX A(2,2),B(2,2),C,D
      C=A(1,1)*A(2,2)-A(1,2)*A(2,1)
      IF((AIMAG(C).EQ.0.).AND.(REAL(C).EQ.0.)) GO TO 10
      D=A(2,2)/C
      B(1,2)=(-A(1,2))/C
      B(2,1)=(-A(2,1))/C
      B(2,2)=A(1,1)/C
      B(1,1)=D
      RETURN
   10 IE=1
      PRINT*,"SINGULAR MATRIX"
      RETURN
      END
```

```
      SUBROUTINE PALROS (P,QL,QU,N,NRS,FNEW,KSEAR,IGOTO)
      SUBROUTINE FOR FINDING THE MINIMUM OF A FUNCTION, USING
C     ROSENBROCK'S METHOD MODIFIED BY PALMER. (SEE M. S. THESIS,
C     D. OLMSTEAD, U OF I, 1972.)
C     P-IS THE PARAMETER VECTOR.
C     QL AND QU ARE THE LOWER AND UPPER BOUND VECTORS
C     N-IS THE NUMBER OF PARAMETERS
C     NRS- IS THE NUMBER OF RANDOM STEPS TO TAKE TO CONFIRM MINIMUM
C     FNEW-(FIRST TIME)-PARAMETER ERROR
C     FNEW-(OTHER)-FUNCTION VALUE
C     KSEAR-(FIRST TIME)-NUMBER OF FUNCTION REQUESTS ALLOWED
C     KSEAR-(OTHER)-SEARCH TYPE.
C                   =0  QUADRATIC ABOUT MINIMUM
C                   =1  QUADRATIC ABOUT FIRST THREE POINTS
C     IGOTO IS NEGATIVE FIRST TIME, THEN RETURNED AND MUST NOT BE
C     CHANGED.
C     IGOTO=1  COMPUTE FNEW, FIRST REQUEST.
C          =2  COMPUTE FNEW, NOT FIRST REQUEST.
C          =3  COMPUTE FNEW, QUADRATIC MINIMUM.
C          =4  ALL DONE, MINIMUM IN P.
C          =5  FUNCTION REQUESTS EXCEEDED, P CONTAINS NEXT VARIABLE SET.
      DIMENSION Q(16),QL(16),QU(16),QH(16),P(16),D(16),A(16,16),B(16,16)
C     INITIALIZATION OF PROCEDURE
      IF (IGOTO) 1,4,4
    1 DDS= FNEW*FNEW
C     DDS IS MINIMUM ACCEPTABLE VARIA[illegible] FOR COMPARISON AFTER 61 IF
C     DESIRED
      DO 3 I=1,N
      Q(I)=P(I)
      DO 2 J=1,N
    2 A(I,J)=0.0
      A(I,I)=1.0
      QH(I)=QU(I)-QL(I)
    3 B(I,1)=SQRT (DDS)*QH(I)
```

```
      KOUNT=1,NRS=KOUNT
      KNTIN=KSEAR
      F1=50.0
      F2=5.0
      XN=N
      MRS=0
      FE=2.0*XN*FNEW*FNEW
      IT=0
      IGOTO=1
      STEP=0.01
      SS=STEP
      CALL RANSET(1)
      RETURN
    4 GO TO (6,7,8,5,5), IGOTO
    5 PRINT 103
      IF (IGOTO-4) 1010,1011,1010
 1010 PRINT 101, IGOTO
 1011 CONTINUE
  101 FORMAT(17H0 INVALID IGOTO  ,I15)
  103 FORMAT(30H0 WE WERE DONE LAST TIME        )
      STOP
    6 PRINT 10
      PRINT 11
      PRINT 12,(Q(I),I=1,N)
      PRINT 13, FNEW
      PRINT 14
      PRINT 12,(QL(I),I=1,N)
      PRINT 15
      PRINT 12,(QU(I),I=1,N)
      PRINT 17
      PRINT 12,(B(I,1),I=1,N)
      PRINT 16
   11 FORMAT(24H INITIAL PARAMETER VALUE)
   10 FORMAT(29H-********** PALROS **********)
   12 FORMAT(7E15.5)
   13 FORMAT(26H INITIAL FUNCTION VALUE F=,E20.12)
```

```
   14 FORMAT(16H LOWER BOUNDS-QL)
   15 FORMAT(16H UPPER BOUNDS-QU)
   16 FORMAT(26H -----------------------    )
   17 FORMAT(41H FINAL ACCURACY OF EACH PARAMETER WILL BE)
C        BASIC ITERATION STEP BEGINS HERE.
C        THIS IS A DO 58 J=1,N WHICH WE MUST ENTER FOR EACH J WITH A NEW
C        VALUE OF F.
   30 J=1
  301 CONTINUE
      FMIN=FNEW
      FMID=FMIN
      DNEW=STEP
      FOLD=FMID
      DMID=0.0
      KK=1
      NSEAR=1
      IGOTO=2
  300 DO 35 I=1,N
      P(I)=Q(I)+A(I,J)*DNEW*QH(I)
C        RETAIN BOUNDS
      IF (P(I)-QL(I)) 33,32,32
   33 P(I)=QL(I)
   32 IF (QU(I)-P(I)) 34,35,35
   34 P(I)=QU(I)
   35 CONTINUE
      KOUNT=KOUNT+1
      NSEAR=NSEAR+KSEAR
      RETURN
    7 IF (FNEW-FMID) 36,37,37
   36 DMIN=DNEW
      FMIN=FNEW
      IGOTO=2
      IF (NSEAR-3) 40,45,45
   40 KK=0
      DOLD=DMID
      DMID=DNEW
```

```
      FOLD=FMID
      FMID=FNEW
      IF (ABS (DNEW)-.01) 41,42,42
   41 DNEW=DNEW*E1
      GO TO 300
   42 DNEW=DNEW*E2
      GO TO 300
   37 DMIN=DMID
      FMIN=FMID
      IGOTO=2
   43 IF (KK) 45,45,44
   44 KK=-1
      FOLD=FNEW
      DOLD=DNEW
      DNEW=-DNEW
      GO TO 300
   45 DELA=FOLD*(DMID-DNEW)+FMID*(DNEW-DOLD)+FNEW*(DOLD-DMID)
      DDEN=(DOLD-DMID)*(DOLD-DNEW)*(DMID-DNEW)
      SECDER=2.0*DELA/DDEN
C     CHECK--MIN OR MAX
      IF (SECDER) 47,47,46
   46 DELB=FOLD*(DMID**2-DNEW**2)+FMID*(DNEW**2-DOLD**2)+FNEW*(DOLD**2-
     1DMID**2)
      D(J)=DELB/(DELA+DELA)
      DNEW=D(J)
      IGOTO=3
      GO TO 300
    8 IF (FNEW-FMIN) 52,52,47
   47 D(J)=DMIN
      FNEW=FMIN
      DO 51 I=1,N
      Q(I)=Q(I)+DMIN*A(I,J)*QH(I)
      IF (Q(I)-QL(I)) 48,49,49
   48 Q(I)=QL(I)
   49 IF (QU(I)-Q(I)) 50,51,51
   50 Q(I)=QU(I)
```

```
   51 CONTINUE
      GO TO 54
   52 FMIN=FNEW
      DO 53 I=1,N
   53 Q(I)=P(I)
   54 CONTINUE
      J=J+1
      IF (J-N) 301,301,60
   60 DD=0.0
      IF (KOUNT-KNTIN) 1999,1999,1998
 1998 IGOTO=5
      GO TO 98
 1999 CONTINUE
      DO 61 I=1,N
   61 DD=DD+D(I)**2
C     CHANGE STEP ACCORDING TO DD.
      U=SQRT (DD)
      V=STEP/10.0
      IF (U-V*10.0) 59,56,56
   56 V=10.0*STEP
      IF (U-V) 58,58,57
   57 IF(V-0.01) 59,59,58
   59 STEP=V
   58 CONTINUE
      IF (DD-DDS) 90,90,62
   62 MRS=0
   65 IF (N-1) 66,80,66
C  ORDER COLUMNS OF A
   66 K=N
      DO 69 J=1,N
      IF (D(J)) 69,67,69
   67 DO 68 I=1,N
   68 B(I,K)=A(I,J)
      P(K)=0.0
      K=K-1
   69 CONTINUE
```

```
      KK=K
      DO 703 L=1,KK
      K=1
      DO 701 I=1,N
      IF (ABS (D(K))-ABS (D(I))) 700,701,701
  700 K=I
  701 CONTINUE
      P(L)=D(K)
      D(K)=0.0
      DO 702 I=1,N
  702 B(I,L)=A(I,K)
  703 CONTINUE
      DO 705 K=1,N
      D(K)=P(K)
      DO 705 I=1,N
  705 A(I,K)=B(I,K)
C     ORTHONORMALIZE COLUMNS OF MATRIX A, BY PALMER@S METHOD, SUCH THAT
C     COLUMN 1 CORRESPONDS TO DIRECTION OF MAXIMUM CHANGE
   70 DO 71 I=1,N
   71 B(I,N)=A(I,N)*D(N)
      P(N)=D(N)**2
      DO 72 L=2,N
      K=N-L+1
      M=K+1
      P(K)=P(M)+D(K)**2
      DO 72 I=1,N
   72 B(I,K)=B(I,M)+A(I,K)*D(K)
      DO 75 J=2,N
      K=N-J+2
      M=K-1
      DIV=SQRT(P(K)*P(M))
      IF (DIV) 73,75,73
   73 DO 74 I=1,N
   74 A(I,K)=(D(M)*B(I,K)-A(I,M)*P(K))/DIV
   75 CONTINUE
      DIV=SQRT(P(1))
```

```
      IF (DIV) 76,90,76
   76 DO 77 I=1,N
   77 A(I,1)=B(I,1)/DIV
C     ITERATION IS COMPLETE
   80 IT=IT+1
      IF (1-MRS) 30,83,83
   83 CONTINUE
      PRINT 81, IT,FMIN,KOUNT
   81 FORMAT(14H ITERATION NO ,I5,3X,4H F =,E15.6,9HFCN REF =,I6)
      PRINT 82, (Q(I),I=1,N)
   82 FORMAT(5H X = ,7E15.5/5X,7E15.5)
      PRINT 15
      GO TO 30
   90 IF (N-1) 9000,95,9000
C     TRY RANDOM DIRECTIONS FOR POSSIBLE IMPROVEMENT
 9000 IF (MRS-NRS) 91,95,95
   91 MRS=MRS+1
      STEP=SS
      IF (MRS-1) 92,92,93
   92 DO 921 I=1,N
      DO 920 J=1,N
  920 A(I,J)=0.0
  921 A(I,I)=1.0
      PRINT 925
  925 FORMAT(11H AXIS RESET)
      GO TO 80
   93 DO 94 I=1,N
   94 D(I)=RANF(IRAN)
      NN=MRS-1
      PRINT 940, NN
  940 FORMAT(15H RANDOM ROTATION,I5)
      GO TO 65
C     PROBLEM TERMINATES
   95 PRINT 96
   96 FORMAT(13H  PALROS TERMINATED)
      IGOTO=4
```

```
98 DO 97 I=1,N
97 P(I)=Q(I)
   RETUPN
   END
```

Vita

William F. Duke was born on 19 February 1940 in Alexandria, Louisiana. He graduated from high school at Angleton, Texas in 1959 and attended one year of college at Texas Technological University in Lubbock, Texas. In 1960 he enlisted in the USAF. Until 1968 he served as a MG-10/Automatic Weapons Systems Mechanic assigned to ADC and the Alaskan Air Command. In 1968 he was accepted into the A.E.C.P. program and studied at Oklahoma State University at Stillwater, Oklahoma, where in 1970 he received the degree of Bachelor of Science in Electrical Engineering. He attended O.T.S. and received a commission in the USAF in late 1970. After nine months' training in the Communication Officers' school at Biloxi, Mississippi he was assigned to Bergstrom A.F.B., Austin, Texas, where he held jobs as Chief-of-Maintenance for CEM in 702nd TAS Squadron, Communications Operations and Planning Officer for the 71st TAS Group, and Chief-of-Maintenance of the 712th DAS Center. In July 1974 he received a resident assignment to Air Force Institute of Technology, School of Engineering to follow a course of study leading to a Master's Degree in Electronic Engineering.

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UNCLASSIFIED

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered) (14) AFIT-GE/EE/77-5

REPORT DOCUMENTATION PAGE		READ INSTRUCTIONS BEFORE COMPLETING FORM
1. REPORT NUMBER GE/EE/77-5	2. GOVT ACCESSION NO.	3. RECIPIENT'S CATALOG NUMBER (9) Master's thesis
4. TITLE (and Subtitle) (6) PARAMETER INDEPENDENT DESIGN UTILIZING SCATTERING PARAMETERS.		5. TYPE OF REPORT & PERIOD COVERED MS Thesis
		6. PERFORMING ORG. REPORT NUMBER
7. AUTHOR(s) (10) WILLIAM F. DUKE		8. CONTRACT OR GRANT NUMBER(s)
9. PERFORMING ORGANIZATION NAME AND ADDRESS AIR FORCE INSTITUTE OF TECHNOLOGY (AFIT/EN) WRIGHT-PATTERSON AFB, OHIO 45433		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS
11. CONTROLLING OFFICE NAME AND ADDRESS **AFAL/WRP** **Air Force Avionics Laboratory** **Wright-Patterson AFB OH 45433**		12. REPORT DATE (11) JUNE 1977
		13. NUMBER OF PAGES ~~108~~ (12) 109p.
14. MONITORING AGENCY NAME & ADDRESS(if different from Controlling Office)		15. SECURITY CLASS. (of this report) UNCLASSIFIED
		15a. DECLASSIFICATION/DOWNGRADING SCHEDULE N/A

16. DISTRIBUTION STATEMENT (of this Report)

APPROVED FOR PUBLIC RELEASE; DISTRIBUTION UNLIMITED

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

18. SUPPLEMENTARY NOTES

APPROVED FOR PUBLIC RELEASE; IAW AFR 190-17

Jerral F. Guess

JERRAL F. GUESS, Capt, USAF
DIRECTOR OF INFORMATION

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

Parameter Independent Design, Scattering Parameters, Computer Aided Design, Optimization, Amplifier Design, Microwave Network Design, Parameter Dependence,

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

A parameter independence factor for two-part networks is defined and a method for its calculation using finite differences is shown. An approach to two-part scattering-parameter circuit design using computer optimization techniques is developed and illustrations are presented to demonstrate the utilization of a digital computer for implementing this approach. The versatility of the approach is clarified by demonstrating how both standard network design criteria and parameter independent network design requirements are specified and met. This design technique has direct Air Force application in

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the areas of microwave network design and Electronic Warfare with particular emphasis on the independence of the network parameters with respect to device parameters.

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